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DESIGN AND IMPLEMENTATION  
OF  
PROJECT 7140 WIDE BAND HF COMMUNICATION TEST FACILITY

B. D. PERRY

July 1981

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## ABSTRACT

The primary objective of project 7140 is to prove the feasibility of the concept of HF communications - using signals having bandwidths on the order of one megahertz - through a series of experiments employing a flexible test facility designed and built for this purpose. The test facility includes a transmitter, equalized receiver, and interference measurement and other support equipments. This report concentrates on the design and realization of the test facility. Descriptions of the wideband HF channel and the planned experimental program are also included.

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## SECTION I

### INTRODUCTION

In this section the report is summarized and its purpose and background are discussed.

#### 1.1 Summary

In the late 1960's, an Air Force ionospheric measurement program conducted by MITRE proved that signals whose bandwidths were ten percent or more of the carrier frequency could be transmitted via HF sky-wave paths provided an appropriate wideband channel equalizer could be realized.<sup>(1,2)</sup> In this current program, such a programmable transversal filter equalizer has been built and installed in a one-way wideband HF sky-wave communication test facility. The equalizer's key parameters are a bandwidth (W) of 1024 kHz, an impulse response duration (T) of 125  $\mu$ s, and a dynamic range of about 50 dB. These parameters are commensurate with reasonable extrapolations of the data taken during the earlier program. The weighting coefficients required to program the equalizer as an inverse filter are derived from probe signal measurements of the transfer function of the ionospheric path. Two phase-coherent, programmable frequency synthesizers are used for probe and fast-frequency-hop spread spectrum signal generation and demodulation. The test facility includes antennas, a wideband transmitter and receiver for the 4- to 30-MHz HF band, and the necessary supporting and peripheral circuitry and equipments. In addition, both terminals of the test facility can be configured for simultaneous interference spectral analysis within the wide bandwidth.

The test facility will be used in a proof-of-concept demonstration of wideband HF communications and for establishing a data base which can be used for future wideband HF modem designs. Data taking and data analysis will emphasize topics such as anti-jamming (AJ), low-probability-of-intercept (LPI) communication, and the effects of interference.

Wideband HF will provide an alternative to satellite and tropospheric-scatter communications and will permit the use of spread spectrum communications with the attendant features of AJ protection and/or LPI communications. In addition, the multipath modes can often be resolved, thereby reducing fading and permitting the option of either increasing the circuit reliability or reducing the transmitted power level.

## 1.2 Purpose

The purpose of this report is to describe the design and implementation of a wideband HF communications test facility under Project 7140: "Wideband HF Technology." This test facility is capable of measuring the characteristics of a wideband HF channel, transmitting spread spectrum signals one-way over that channel, and measuring the noise and interference present at both the transmit and receive terminals of the channel.

## 1.3 Background

Communications links in the 3- to 30-MHz HF band are important to many classes of users, both civilian and military. Civilian applications include international "shortwave" broadcasting services, radio telegraphy, ship-to-shore communications, and transoceanic air traffic control. All three military services have relied and continue to rely on HF links for both short-haul tactical communications

and longer-distance strategic links. The principal advantage of HF sky-wave communications is its relatively low cost for beyond-line-of-sight applications. Its many disadvantages, however, include limited channel bandwidths and various degradations caused by the vagaries of the ionosphere (and earth reflections for multi-hop links).

The earlier MITRE work<sup>(1,2)</sup> involved the measurement of the characteristics of oblique HF paths over bandwidths of up to 4.5 MHz. This work led to the discovery that the path parameters varied quite slowly and to the consequent conclusion that path equalization would be feasible once the requisite technology became available. With the advent of large-scale and very-large-scale-integrated (LSI and VLSI) signal processing devices path equalizers for bandwidths of 1 MHz or greater can now be realized.

Project 7140 began on 1 October 1978. During the first year, alternative equalizer design approaches were considered, one was chosen, a flexible test facility including the chosen equalizer was designed, and implementation began. During this the second year, the implementation has been completed. In the process of completion certain portions of the test facility were redesigned to overcome inadequacies in the original design and make use of resources not previously available.

#### 1.4 Outline

In section II the characteristics of wideband HF channels are summarized. These characteristics are then used as a basis for the

test facility design approach described in section III and test facility implementation described in section IV. The planned experimental program using the test facility is described in section V.



## SECTION II

### WIDEBAND HF CHANNEL CHARACTERISTICS

In this section the existing information on wideband HF propagation channel characteristics is summarized to provide a reference for the design approach described in section III.

#### 2.1 Ionograms

The most common technique for measuring the characteristics of the ionospheric propagation paths that exist between two points (i.e. the terminals of a radio communications circuit) is to use oblique ionospheric sounding equipment and generate "oblique ionograms." An ionogram is an intensity-modulated display of the time of arrival of the transmitted pulse (or its equivalent as in the case of a "chirp sounder") as a function of frequency across the entire HF band. A typical ionogram is shown in Figure 1. It exhibits one-, two-, and three-hop modes. The primary mode of propagation is the lowest, or one-hop, mode. As can be seen, it is dispersive, with its time of arrival varying from 1.0 ms at 16 MHz to approximately 1.2 ms at the "nose" of the trace (commonly referred to as the maximum usable frequency or MUF) at 22.5 MHz. Between 20 MHz and the MUF, the upper ray (sometimes called the Pederson ray) is visible. Thus there remains a "window" between about 15 and 20 MHz where only a single mode propagates and where intermode multipath fading will not occur.

There are two other features of importance displayed in this ionogram. The vertical traces are caused by the strongest of the many narrowband emitters which were interfering with the reception of the sounding signal. This interference is usually most predominant in the international AM broadcast bands. The somewhat periodic

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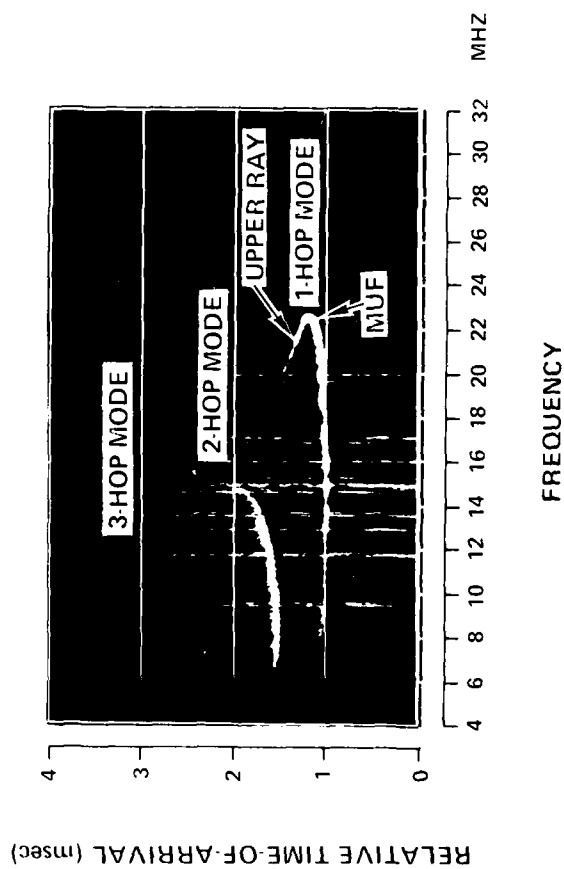


Figure 1. OBLIQUE IONOGRAM

intensity modulation, which is especially apparent on the one-hop trace, is due to the Faraday rotation effect. The transmitted linearly polarized wave is constituted of two components - a left-hand circular component and a right-hand circular component. These components are affected differently by the earth's magnetic field. Consequently, they arrive at the receiver with a small difference in time of arrival. (Typically the difference is a few microseconds on oblique paths.) This effect leads to a slow fading rate at any particular frequency as the received wave rotates in polarization with respect to the receiving antenna polarization. In addition this effect leads to a frequency-dependent interference pattern which, as can be seen, exhibits minima about every few hundred kHz.\*

## 2.2 Transfer Functions

Another way of describing the sky-wave propagation between two points is by assuming it can be modeled by a linear time-invariant network and its associated transfer function. The assumption of linearity is generally valid, except for the rarely observed Luxemburg effect\*\*. The assumption of time invariance is justified for wide-band HF propagation by the earlier MITRE work<sup>(1,2)</sup> which showed time invariance to typically be 10 seconds (but up to at least 30 seconds during the daytime and as little as one second during dawn-dusk transitions).

---

\*For a more complete description of Faraday rotation and its effect on oblique path characteristics, see reference 3.

\*\*This effect is ionospheric cross-modulation between two transmissions on different carrier frequencies. It is only observed when high-power transmitters are involved such as the venerable Radio Luxemburg.

### 2.2.1 Full HF Band

We define the transfer function as  $C(\omega) = E(\omega)e^{jD(\omega)}$  and show in Figure 2 a simplified sketch of the magnitude  $E(\omega)$  and the negative phase derivative or group delay  $T_g(\omega)$  as a function of frequency that one would expect for the path associated with the ionogram of Figure 1. (Group delay is shown rather than phase because it is more closely related to the ionogram.) For those frequencies where more than one propagation mode exists (i.e., one-hop plus two-hop or upper plus lower ray) the magnitude term varies periodically with maxima spaced by the reciprocal of the delay difference between the modes. In addition Faraday rotation causes a fluctuation in the magnitude term, as described in section 2.1 above, for each mode. This fluctuation is not generally observable for multi-hop modes, however, because the differential delay associated with this effect is less than the smearing in time of arrival due to the nonspecular earth reflection(s).

### 2.2.2 Band-limited

The typical wideband channel will be on the order of one megahertz in extent (see sections 2.3 and 3.1). In many cases it will be possible to select a one-megahertz band where only the one-hop mode propagates and where the upper ray is at least 20 dB down with respect to the lower ray. (In Figures 1 and 2 this would be the region between 15 and about 21.5 MHz.) This simplifies the channel equalizer and leads to the design incorporated in the test facility which is the subject of this report. In those cases where significant energy propagates by more than one (widely separated) mode (within the band chosen), this simplified equalizer is still useful for spread spectrum communications as will be described in 3.1.

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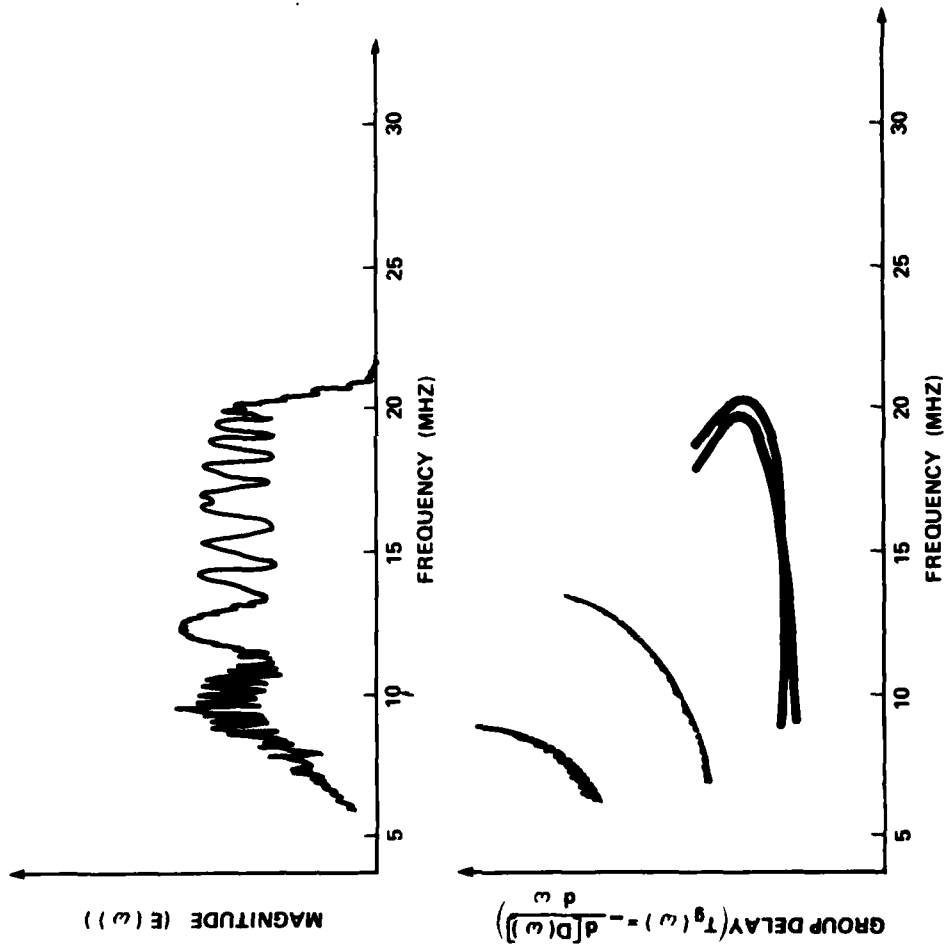


Figure 2: SKETCH OF TRANSFER FUNCTION  $[C(\omega) = E(\omega) e^{jD(\omega)}]$  FOR FULL HF BAND

### 2.3 Measured Parameters

The earlier MITRE program<sup>(1,2)</sup> provided a somewhat limited but nevertheless useful data base with respect to the characteristics of wideband HF channels. The data base was limited to one-hop cases as described in the previous section and was confined to the temperate latitudes where auroral effects are not present. The data base is particularly useful in that from it we can confidently extrapolate to a set of equalizer parameters for the design incorporated in this test facility. Figure 3 is a sketch of the typical band-limited transfer function represented by this data base. The bandwidth ( $W$ ) is that of (1) the signal transmitted to measure or "probe" the channel transfer function and (2) the inverse-filter equalizer used to compensate for the channel. Within this band,  $\Delta E$  is the peak-to-peak magnitude variation,  $\Delta T_g$  is the maximum group delay difference or the "dispersion," and  $\Delta \tau$  is the average delay difference between the so-called "ordinary" and "extraordinary" ray components due to the magneto-ionic splitting (i.e. Faraday rotation).

It should be noted that  $\Delta T_g$  in a given band  $W$  (say one megahertz) increases as the path length is reduced and as the carrier frequency ( $f_o$ ) is also reduced so as to stay below the MUF. To minimize this dependence of  $\Delta T_g$  on  $f_o$  the restriction that  $W \approx 0.1f_o$  is applied. We can then summarize the one-hop data base of the earlier MITRE work as follows:\*

For  $W \approx 0.1f_o$ :

- $\Delta E$  is observed to be "less than 30 dB most of the time".
- $\Delta T_g$  is observed to be "a few tens of microseconds most of the time," but over 100  $\mu s$  at or near dawn and dusk.
- $\Delta \tau$  is observed to be in the order of "a few microseconds".

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\*For very short paths where the propagation can not be characterized as oblique,  $\Delta T_g$  is somewhat larger and the data as summarized here does not necessarily apply.

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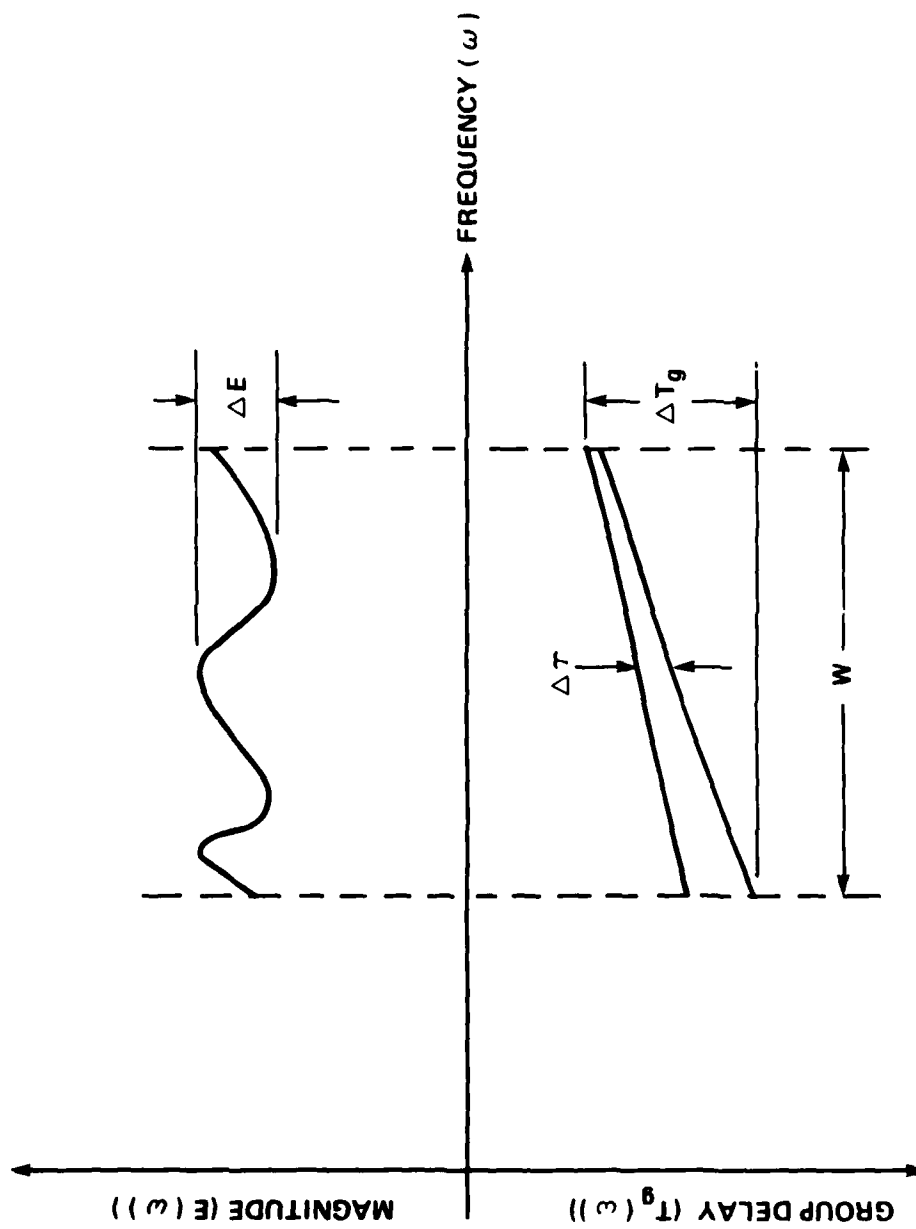


Figure 3: TYPICAL BAND-LIMITED TRANSFER FUNCTION FOR SINGLE-MODE OBLIQUE PROPAGATION

These observed parameters were typically time invariant over 10 seconds.

#### 2.4 Noise and Interference

The terms noise and interference have somewhat unique connotations in the HF band. The predominant noise is external to the receiver and is due to both man-made (ignitions, power lines, machinery, etc.) and natural (thunderstorms, galactic sources) causes. This external noise is best distinguished from interference in that it is broadband. The noise bandwidths involved do not always encompass the entire HF band but are at least greater than the bandwidths considered under "wideband HF communications." The methods used to calculate and otherwise deal with noise are therefore no different than those used in more traditional narrowband HF communications.

HF band interference is narrowband, however, since it consists of the emissions from various transmitters operating in their assigned narrowband channels. The methods used to deal with interference in the wideband HF case are different from those used in the traditional narrowband situations. The narrowband user attempts to avoid the interference by selecting a quiet channel. The wideband user can never expect to totally avoid the interference (although he can certainly attempt to avoid the more crowded frequency bands, at least in the daytime). The wideband user can, however, gain an advantage over the interference through signal processing in that the wideband receiver will be mismatched to the typical narrowband waveform. In addition, the wideband user might choose to (1) modify the transmitted signal so as to avoid transmitting in time and/or frequency slots which are occupied by others or (2) use noise whitening signal processing techniques in the receiver.



### SECTION III

#### DESIGN APPROACH

In this section the approach taken to the design of the test facility is discussed. This approach is to:

- 1) design a wideband equalizer which will compensate for the channel in the receiver by operating as a programmable filter whose transfer function is the inverse of that of the channel,
- 2) provide a means of measuring the channel transfer function and from it calculate the equalizer filter coefficients,
- 3) transmit wideband "spread-spectrum-like" signals over the equalized channel, and
- 4) provide a means of measuring the characteristics of the interference within the wideband channel.

Thus the facility can be used to expand the data base of the earlier MITRE work, prove the concept of wideband HF communications, and determine what can and/or should be done to mitigate the effects of interference. In essence the facility will measure and deal with the vagaries of the channel transfer function and measure but not directly deal with the effects of interference.

#### 3.1 Selection of Equalizer Parameters

The need for a channel equalizer arises when one attempts to use a bandwidth greater than the correlation bandwidth of the ionosphere. For multi-mode situations, this bandwidth is as low as 0.3 kHz. For single, one-hop mode situations, this bandwidth is on the order of 50 to 100 kHz.\* These facts together with our previous experience<sup>(1,2)</sup>

\*This value is an average over time of day and geography. Lower values are to be expected at dawn/dusk and in auroral regions. Conversely, values as large as several hundred kHz have been observed in temperate latitude, daytime situations.

indicated that an equalizer bandwidth on the order of one megahertz or ten percent of the carrier frequency (whichever is smaller) would be appropriate and could be realized using present-day circuit technology. For design convenience  $W = 1024$  kHz has been selected with an option to use one half the bandwidth (512 kHz).

Channel equalization using inverse filtering is not the optimum strategy in all circumstances. For example, conjugate filtering, wherein the adaptive receiver is configured as a matched filter to the convolution of the transmitted waveform and the channel impulse response, is a better strategy for low signal-to-noise applications. However, high signal-to-noise ratios are the rule in HF communications (except where strong interference predominates or where either absorption or multipath cause the signal to suffer a deep fade exceeding the transmitter power reserve designed for this contingency). We therefore provide sufficient transmitter power, expect a high receiver signal-to-noise ratio, and use an inverse filter equalizer.

Other equalization strategies can be implemented by using a method of calculating equalizer coefficients that is different from the method described in section 3.3. Later in the program, the test facility may be modified to accommodate one or more of these strategies.

The second important equalizer parameter is the duration of its impulse response ( $T$ ). In order to correct for dispersion, it is only necessary that  $T$  be equal to or greater than  $\Delta T_g$ . To correct for multipath, however,  $T$  must exceed  $\Delta \tau$  by about an order of magnitude. Even then it will be insufficient in the pathological case wherein:

- 1) the two multipath signals are exactly equal in magnitude,
- 2) there consequently are zeros in the channel transfer function, and thus
- 3) the required  $T$  goes to infinity.

The equalizer is realized as a finite-impulse-response filter and the value of  $T$  chosen is  $125 \mu\text{s}$ . This value satisfies both the requirements of  $T \geq \Delta T_g$  and  $T \gg \Delta \tau$ , at least in the vast majority of expected situations.

The third important parameter is dynamic range. In section 2.3 it is indicated that 30 dB will be sufficient. The chosen value is about 51 dB which allows a more than 20-dB worst-case margin in the presence of strong in-band interfering signals. It is achieved by 8-bit rectangular coordinate processing.

Although the selection of this  $W = 1024 \text{ kHz}$  and  $T = 125 \mu\text{s}$  (and 8-bit processing) drive the equalizer design realization and describe its capability, the equalizer can also be reconfigured for three other arrangements. The following are the four alternatives:

1.  $W = 1024 \text{ kHz}$ ,  $T = 125 \mu\text{s}$ . (128 samples 8 kHz apart in frequency and  $0.977\text{-}\mu\text{s}$  tap spacing.)
2.  $W = 1024 \text{ kHz}$ ,  $T = 62.5 \mu\text{s}$ . (64 samples 16 kHz apart in frequency and  $0.977\text{-}\mu\text{s}$  tap spacing.)
3.  $W = 512 \text{ kHz}$ ,  $T = 125 \mu\text{s}$ . (64 samples 8 kHz apart in frequency and  $1.95\text{-}\mu\text{s}$  tap spacing.)
4.  $W = 512 \text{ kHz}$ ,  $T = 250 \mu\text{s}$ . (128 samples 4 kHz apart in frequency and  $1.95\text{-}\mu\text{s}$  tap spacing.)

These alternatives permit measurements of relative effectiveness. For example, during disturbed ionospheric conditions the fourth alternative would possibly be the best choice, whereas, for very short paths which require low carrier frequencies, the third or fourth might be best. The second and third alternatives relax the equalizer design requirements and permit an evaluation of the usefulness of simpler equalizer designs than the primary design being implemented under this program. (The program also plans to investigate the effectiveness of the full 8-bit dynamic range capability as compared with lesser capabilities down to 1-bit processing.)

### 3.2 Measurement of Channel Transfer Function

Because the channel is band limited to  $W$  and its impulse response is expected to be no greater than  $T$ ,  $2TW$  samples in either the time or frequency domain are sufficient to characterize it. Given  $W = 1024$  kHz and  $T = 125$   $\mu$ s,  $2TW = 256$ . Rather than impulsing the channel, we use frequency domain sampling in order to minimize the peak power radiated for the probe signal and to minimize our interference to other users. Thus we require 128 complex samples every  $1/T = 8$  kHz. This can be instrumented, as shown in Figure 4, as the simultaneous transmission of 128 complex sinusoids 8 kHz apart ( $1/T$  has been defined here as  $\Delta F$ ) from a bank of phase-coherent frequency sources.

Since the channel is not time invariant, however, we must complete the probe transmission in less than the "correlation time" of the ionospheric channel. For the one-second probe transmission time ( $T_p$ ) used, the probe signal spectrum lines broaden to 1 Hz widths, as shown in Figure 5.

A simplification is achieved by time multiplexing the probe frequency-domain samples, providing sufficient power to guarantee a good receiver signal-to-noise ratio for each sample, and completing the 128 transmission segments in one second. The width of each spectral line, which is now 128 Hz, is still much less than  $\Delta F = 8000$  Hz. Figure 6 shows both the time and frequency domain representations of this probe signal.

In order to discriminate against narrowband interference and achieve mode isolation so that the probe signal can be used to measure the ionospheric path transfer function for the desired mode (which for example in Figure 2 is the one-hop mode), each probe segment is coded with 8 kHz of linear frequency modulation (LFM). Without this coding, the typical mode separation of about one millisecond

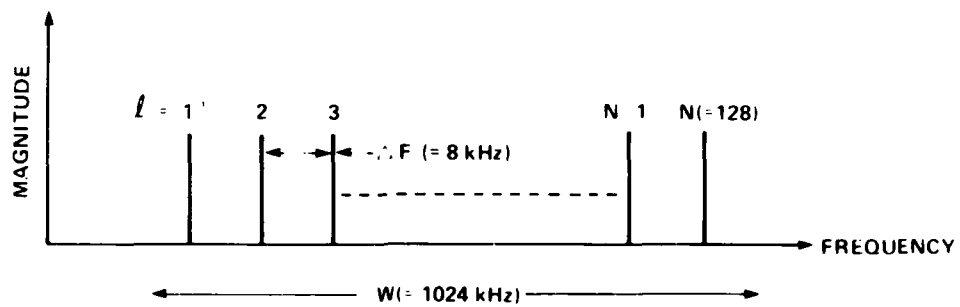


Figure 4: PROBE SIGNAL SPECTRUM; NO TIME TRUNCATION

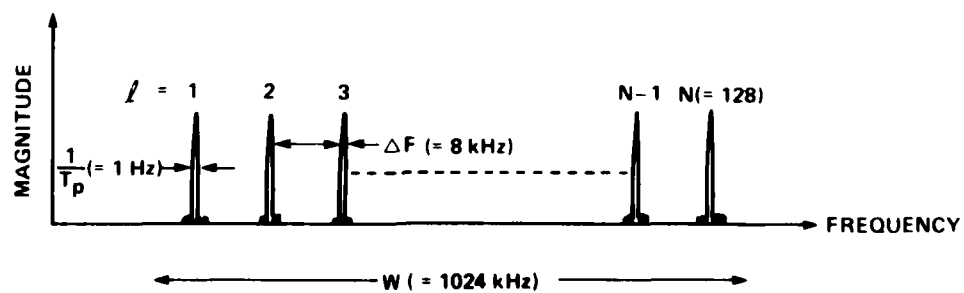
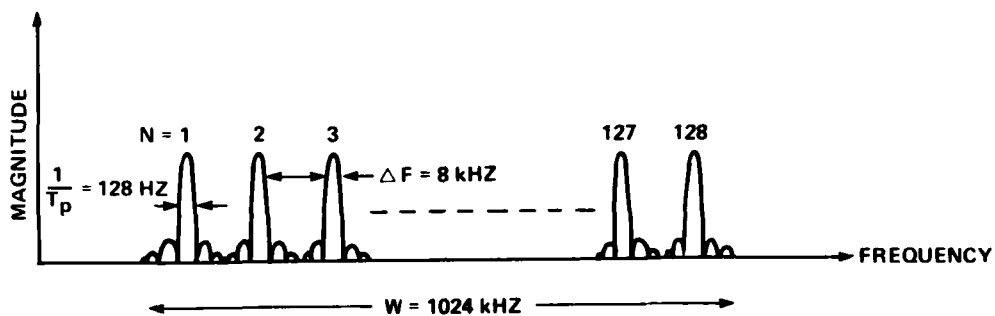
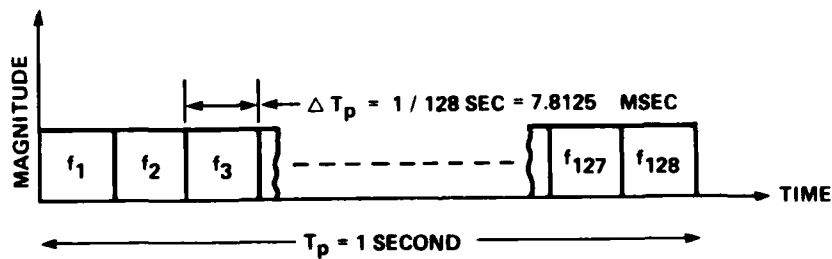


Figure 5: PROBE SIGNAL SPECTRUM; ONE SECOND TIME TRUNCATION



6a: Spectrum



6b: Time Sequence of Probe Signal Segments

Figure 6: PROBE SIGNAL: TIME-MULTIPLEXED TRANSMISSION

is less than the time-domain resolution of the segment of 7.8125 ms.

Figure 7 shows the resulting LFM probe signal. The channel whose bandwidth (W) is 1024 kHz is swept in the frame time ( $T_p$ ) of one second. The sweep consists of 128 contiguous segments. The time resolution of each segment, which is equal to the reciprocal of the segment's 8-kHz bandwidth or 125  $\mu$ s, is sufficient to permit mode separation. In addition there is no need to increase the peak power as would be the case if uncoded probe segments less than one millisecond long were used.

### 3.2.1 Probe Signal Acquisition

In order to permit acquisition of the desired-mode signal, the time of arrival must be known or determined to within a small fraction of the segment duration by searching over the residual time uncertainty. This time can be estimated as follows:

- The uncertainty in time of arrival is the sum of the clock time-of-day errors at the transmit and receive sites plus the uncertainty in the propagation delay over the path being measured.
- Using HF time stations (such as WWV in Ft. Collins Colorado or CHU in Ottawa, Canada) to establish time of day, the total uncertainty is that associated with three HF paths and is in the order of two milliseconds.

Figure 8 shows the method used to search this two millisecond uncertainty. The received LFM probe signal is demodulated by mixing with an LFM local oscillator (LO). As a result, any time difference between the two LFM signals is converted into a frequency difference. The low pass filter (LPF) thus serves as a time window which discriminates against received signals whose time of arrival differs from the start time of the LO frame by more than one half of the window. This window's duration is equal to

$$T_c = \frac{2 f_c T_p}{W} \quad (1)$$

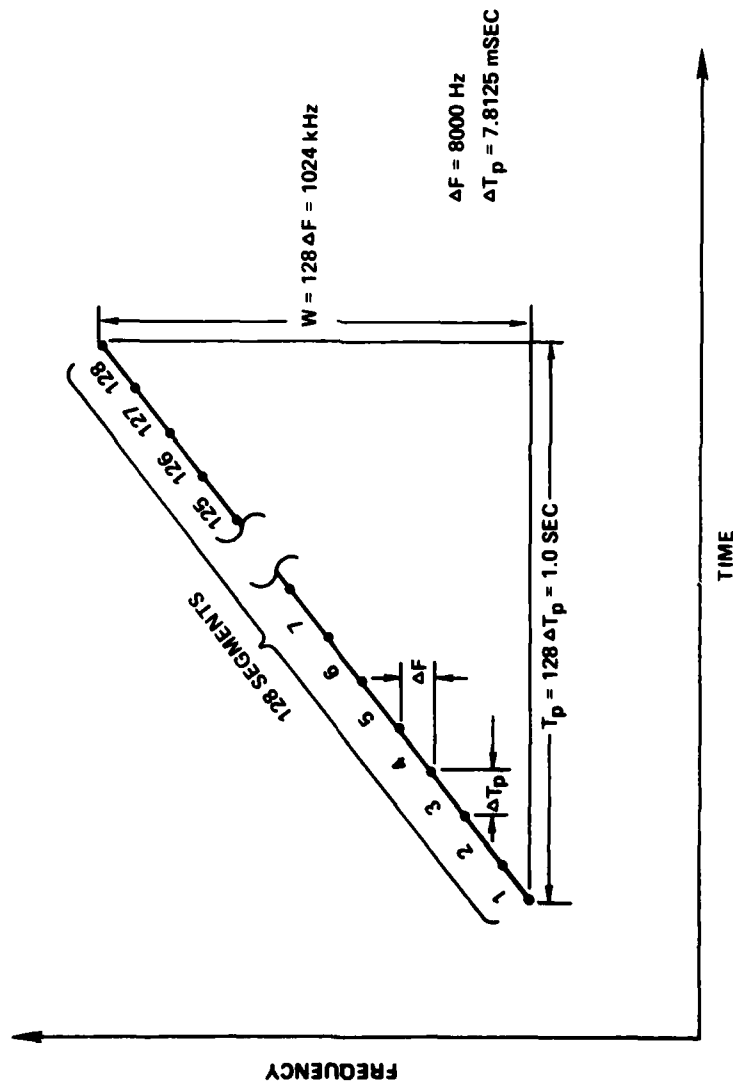


Figure 7: PROBE SIGNAL; LFM CODING ON EACH SEGMENT, NO SCRAMBLING



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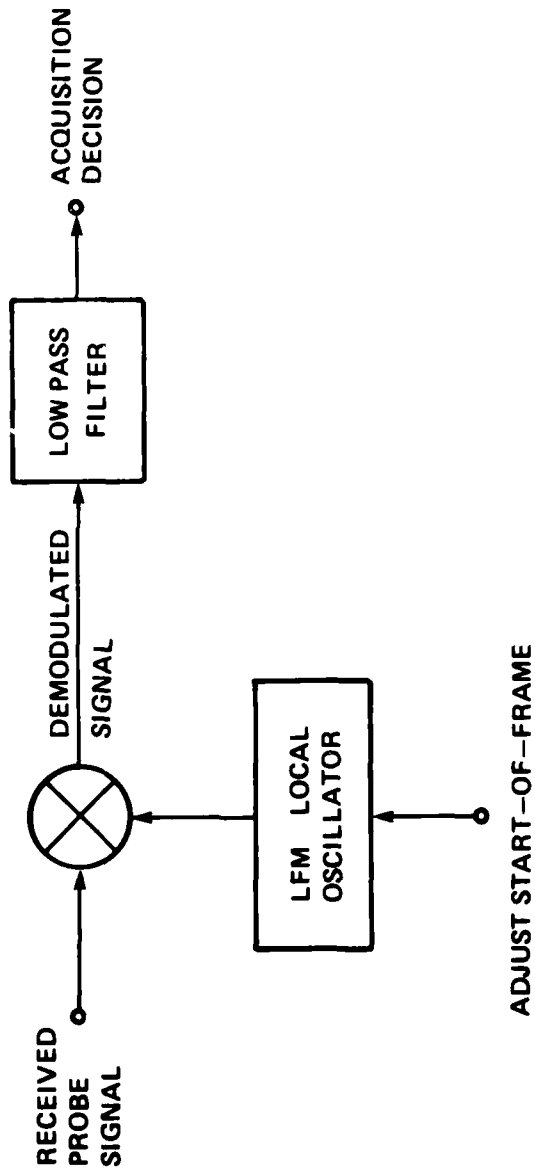


Figure 8: PROBE ACQUISITION CIRCUITRY

where  $f_c$  = LPF bandwidth in Hz,  
 $T_p$  = probe frame time in seconds,  
 and  $W$  = bandwidth in Hz.

The LPF is however not matched to the probe segment. Instead, a lower signal-to-noise ratio is accepted during acquisition in order to reduce the time required to search over the two millisecond uncertainty interval.

In the test facility,  $f_c = 400$  Hz and  $T_p = 1$  second. For  $W = 1024$  kHz, equation (1) yields  $T_c = 781$   $\mu$ s. Since each trial setting of start of frame must be held for  $T_p$  (one second), the total time to acquire the desired signal within  $T_c$  is given by

$$T_{acq} \leq \frac{1 \times 2000}{781} \text{ or about 2.5 seconds.}$$

Once acquired within the window (and assuming a good signal-to-noise ratio), the residual time error will rarely exceed 10% of the window width or about 80  $\mu$ s. Further reduction in the time error can often be achieved until finally limited by the dispersion of the path.

After acquisition the probe signal is time sampled to provide the desired 128 complex samples of the channel transfer function,  $C(\omega)$ . As shown in Figure 9, a second down-converter to baseband is used, but with the demodulating LFM signal applied in quadrature phase so that rectangular coordinate samples  $\text{Re}\{C(\omega)\}_\ell$ , and  $\text{Im}\{C(\omega)\}_\ell$ , are obtained. (The subscript  $\ell$  refers to the  $\ell$ th sample in the set of  $N = 128$ ).

It has been shown that time sampling the received LFM waveform in this way yields the desired frequency-domain samples of  $C(\omega)$ . The proof rests on the fact that the time-bandwidth product of the probe signal is much greater than that of the channel (for a single mode) and is demonstrated by comparing the phase rate of

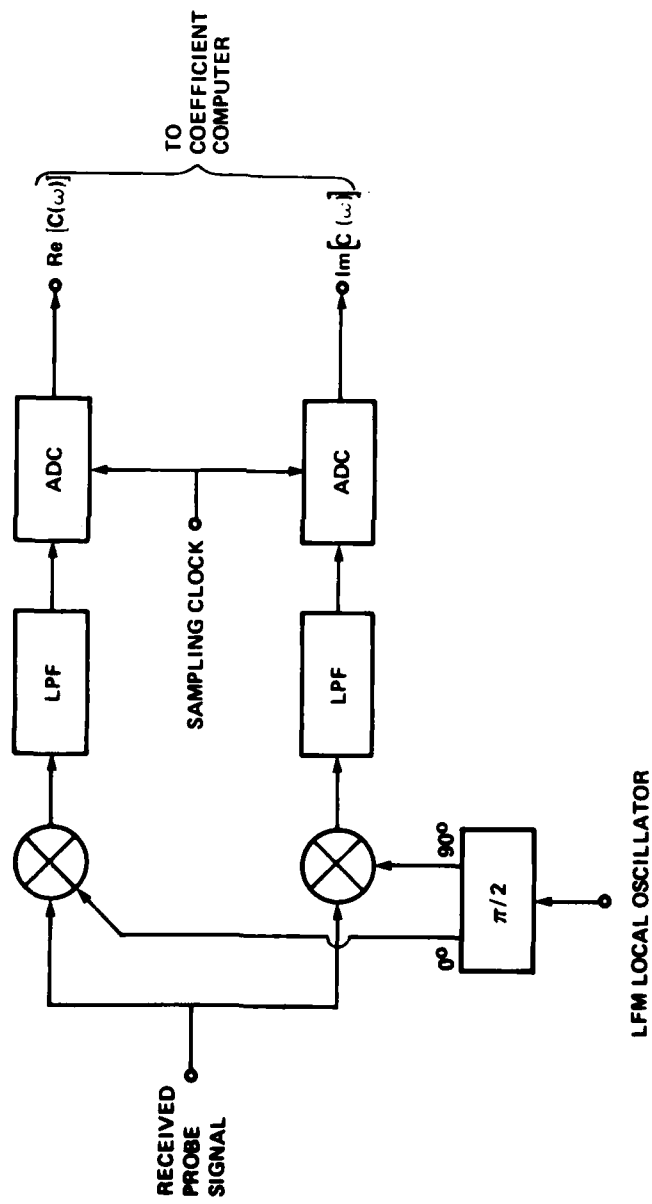


Figure 9: PROBE SAMPLING FOR CHANNEL TRANSFER FUNCTION MEASUREMENT

change versus frequency of the LFM sweep with that of the channel transfer function.

Finally, to minimize annoyance to other users, the N probe segments are transmitted in a scrambled order.

### 3.3 Computation of Equalizer Coefficients (for Inverse Filtering)

Given a band-limited channel whose transfer function is  $C(\omega)$ , any signal  $S(\omega)$  limited to the same band can be transmitted through the channel and recovered without distortion if the channel is followed with an inverse filter whose band-limited transfer function is  $C(\omega)^{-1}$ , as shown in Figure 10. Using polar form representation

$$C(\omega) = E(\omega) e^{jD(\omega)} \quad (2)$$

and therefore

$$C^{-1}(\omega) = \frac{1}{E(\omega)} e^{-jD(\omega)} \quad (3)$$

The typical  $l$ th member of the set constituting  $C(\omega)$  sampled is

$$C_l^{-1} = \frac{1}{E_l} e^{-jD_l} \quad (4)$$

It becomes necessary to compute the equalizer coefficients in rectangular coordinates because the device is most easily realized in this form (see section 4.12). Before discussing the time-domain equalizer actually implemented, we first consider the equalizer as a bank of N filters and associated weighting circuits with an output combiner. Each filter passes one of the N frequency domain samples of  $C(\omega)$ . This type of equalizer is shown in Figure 11.

Here

$$C_l^{-1} = X_l + jY_l \quad (5)$$

where

$$X_l = M_0 \operatorname{Re} \left( \frac{1}{C_l} \right) \text{ and } Y_l = M_0 \operatorname{Im} \left( \frac{1}{C_l} \right)$$

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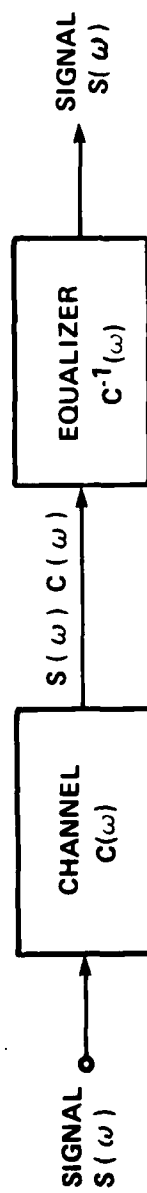


Figure 10: INVERSE FILTER EQUALIZATION

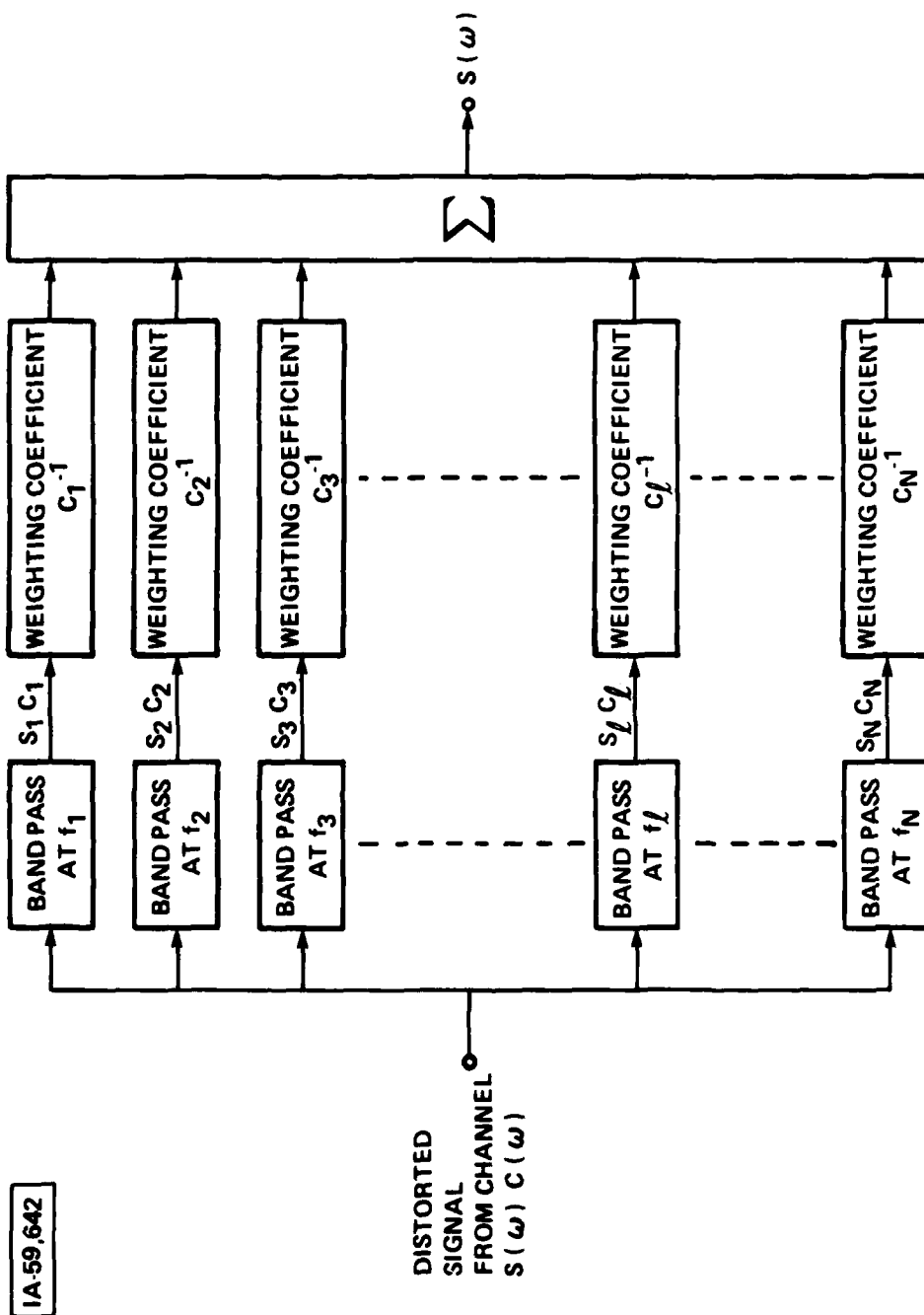


Figure 11: EQUALIZER AS A BANK OF FILTERS

and  $M_0$ , the normalizing factor,  $= \frac{1}{N} \sum_{\ell=1}^N E_{\ell}$

Thus

$$X_{\ell} = \frac{M_0}{E_{\ell}^2} \operatorname{Re} (C_{\ell}) \text{ and } Y_{\ell} = - \frac{M_0}{E_{\ell}^2} \operatorname{Im} (C_{\ell}) \quad (6)$$

and the circuitry for each filter's weighting accomplishes complex multiplication in rectangular coordinates

$$\left\{ \operatorname{Re} (S_{\ell} C_{\ell}) + j \operatorname{Im}(S_{\ell} C_{\ell}) \right\} \cdot \left\{ X_{\ell} + jY_{\ell} \right\}$$

### 3.3.1 Fast Convolution

Fast convolution is the name often applied to a digital filtering technique<sup>(4)</sup> for realizing frequency-domain equalization. It is described in simplified form in Figure 12. The received signal  $g(t)$  is the convolution of the transmitted waveform  $s(t)$  and the band-limited channel impulse response  $c(t)$ .

That is

$$g(t) = s(t) * c(t) \quad (7)$$

The analog signal  $g(t)$  is converted to digital form, transformed by a discrete Fourier transform (DFT) process such as the FFT or the Chirp-Z transform (CZT).<sup>(5)</sup> The DFT operates on N-point blocks of data and provides the frequency-domain constituents of the signal one at a time to a complex multiplier for the appropriate weighting. The inverse DFT then reconstructs N-point blocks of the equalized time-domain waveform,  $s(t)$ . To avoid the circular convolution implicit in the process shown in Figure 12, overlapping blocks of data are padded with zeros, processed in a parallel channel, and summed.

Fast convolution was not selected as the technique for realizing the equalizer in the test facility since, at the time of investigation in late 1978, it offered no advantage over the time-domain equalizer selected. It is mentioned here because (1) it is felt that

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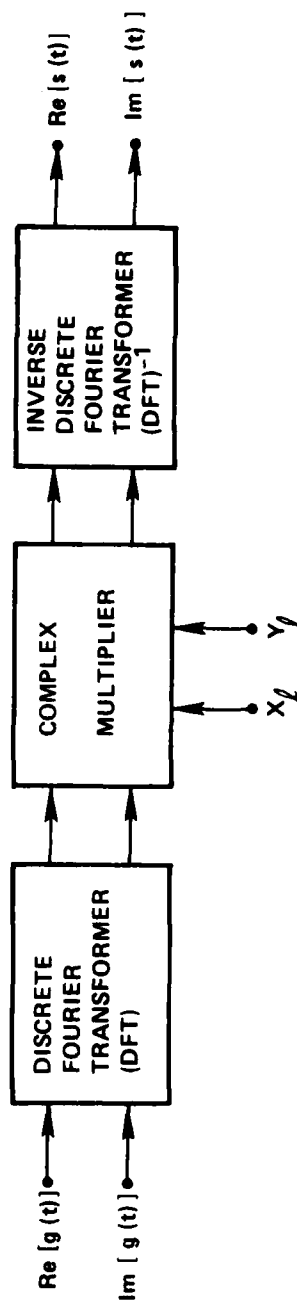


Figure 12: EQUALIZATION IN THE FREQUENCY DOMAIN VIA FAST CONVOLUTION



subsequent advances in VSLI technology in general and in CCD technology in particular have altered the balance somewhat in the direction of fast convolution and (2) wideband HF modems, in contrast to this facility, will need to be designed with a view toward minimum size, weight, and power. Consequently, use of the latest technology may be required.

### 3.4 Transversal Filter Equalizer

In the previous section we described how the probe signal yields the channel transfer function,  $C(\omega)$ , and how this function could be inverted to provide the weighting coefficients for frequency-domain equalization. In this section we describe the test facility equalizer which operates directly on the received waveform  $g(t)$  to restore it to its original form,  $s(t)$ . This equalizer can equally well be described as (1) an inverse filter, (2) a transversal filter, or (3) a deconvolver. It is a band-limited device operating within the band  $W$ . It computes  $h(t)$ , the convolution of  $g(t)$  and the appropriate weighting function  $v(t)$ . Thus we anticipate that

$$h(t) = g(t) * v(t) = s(t), \quad (8)$$

and the device deconvolves  $s(t)$  and  $c(t)$  within the band  $W$ . As can be seen from the diagram of the equalizer shown in Figure 13,

$$h(t) = \sum_{k=1}^N g(t-k\tau) v_k \quad (9)$$

where

$$v_k = A_k + j B_k$$

Since  $N=128$  samples of  $C(\omega)$  are provided by the probe signal, it is convenient to have 128 taps on the tapped delay line.\*

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\*128 is also convenient with respect to the modular design described in section 4.1.2.

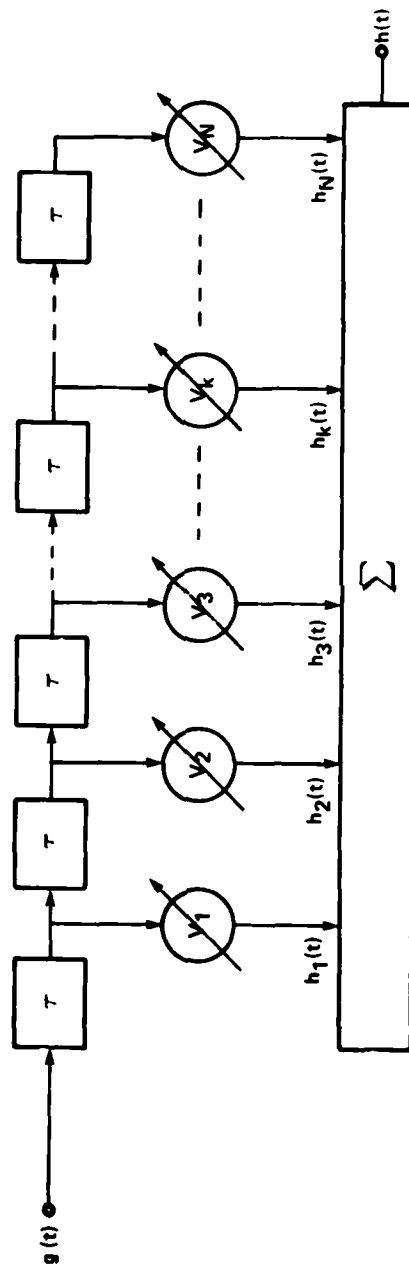


Figure 13: TAPPED DELAY LINE, TRANSVERSAL FILTER, DE-CONVOLVING EQUALIZER

We next compute the equalizer's transfer function to determine the tap weights required for inverse filtering. The band-limited impulse response of the equalizer is seen by inspection to be

$$W \sum_{k=1}^N V_k \frac{\sin \pi W(t-k\tau)}{\pi W(t-k\tau)} \cdot e^{j\omega_0(t-k\tau)} \quad (10)$$

where  $\omega_0$  is the mid-band frequency (the "carrier").

The Fourier transform of this impulse response, which is the transfer function of the equalizer and its associated band-limiting filter, is

$$\sum_{k=1}^N V_k e^{-j\omega\tau} \text{ for } \omega_0 - \frac{W}{2} < \omega < \omega_0 + \frac{W}{2} \quad (11)$$

and zero elsewhere.

But this is the expression for the DFT. Thus the equalizer frequency response is the DFT of the tap weights. Since we aim to implement a filter whose transfer function is  $C(\omega)^{-1}$ , we conclude that the tap weights are the inverse DFT of the coefficients previously calculated in 3.3.

In summary then:

- The probe signal provides the 128 complex samples of  $C(\omega)$ .
- $C(\omega)^{-1}$  is calculated resulting in 128 frequency-domain coefficients. Typically  $C_\ell^{-1} = X_\ell + j Y_\ell$ .
- The 128 complex tap weights are obtained from the 128  $C_\ell^{-1}$ 's via the inverse DFT.

In the event that there is a zero in  $C(\omega)$ , the equalizer transfer function's magnitude should go to infinity at the frequency of the zero amplitude. The phase, however, is indeterminate. In practice, if and when this occurs, an approximation is employed.

In the event of a narrowband interfering signal, which is not negligible with respect to the probe, a corrupted measurement will result. The equalizer will, however, tend to provide a null at this

frequency, probably the best strategy for this case. Later when we have better understanding of interference (see section 3.6), recommendations for more optimum processing strategies can be made.

### 3.5 Spread Spectrum Test Signals

Given an equalized wide bandwidth HF channel, the question arises as to the approach to be taken to make use of this channel. We have chosen spread spectrum signalling in order to develop a data base which is relevant to Air Force needs for AJ and/or LPI communications. The two important classes of spread spectrum waveforms are (1) frequency hopping (FH) and (2) direct sequence pseudo-noise phase modulation (usually called "PN"). Since an FH capability is implicit in the test facility probe signal transmission and reception, we have chosen to implement this waveform first. Later it is planned to add a PN capability (see section 4.7).

#### 3.5.1 FFH Waveform

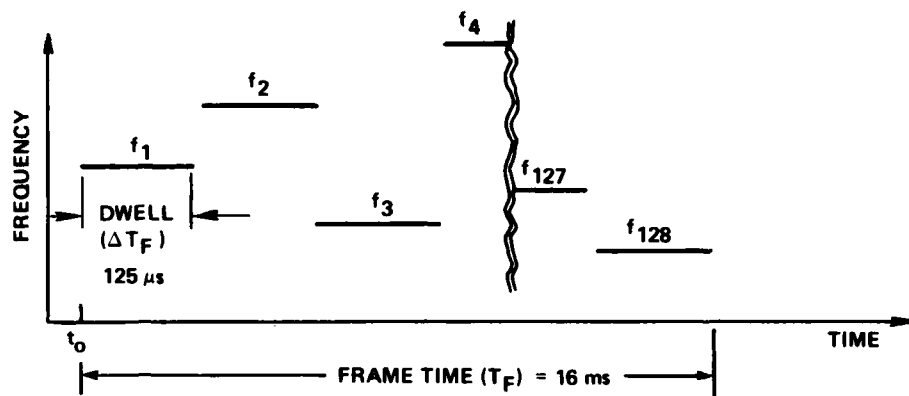
The FH waveform used in the test facility is called the fast frequency hop (FFH) waveform. It is derived from the probe signal by (1) omitting the LFM feature and transmitting at a single frequency per segment (or "dwell"), and (2) reducing the duration of the dwell to less than the usual ionospheric mode path-delay difference so as to permit mode resolution by way of time gating in the receiver. A dwell duration of 125  $\mu$ s is used in the system so as to be compatible with probe and equalizer parameters. This in effect provides a time gate of 125  $\mu$ s. Within this 125  $\mu$ s time window the equalizer compensates for delay and amplitude variations including those due to Faraday rotation.

The FFH waveform is phase-coherent over the 128 dwells which constitute a frame. This frame is 16 ms long and can be used for the transmission of one or more data bits. One bit per frame will result in a 42 dB processing gain and a 62.5-b/s data rate.

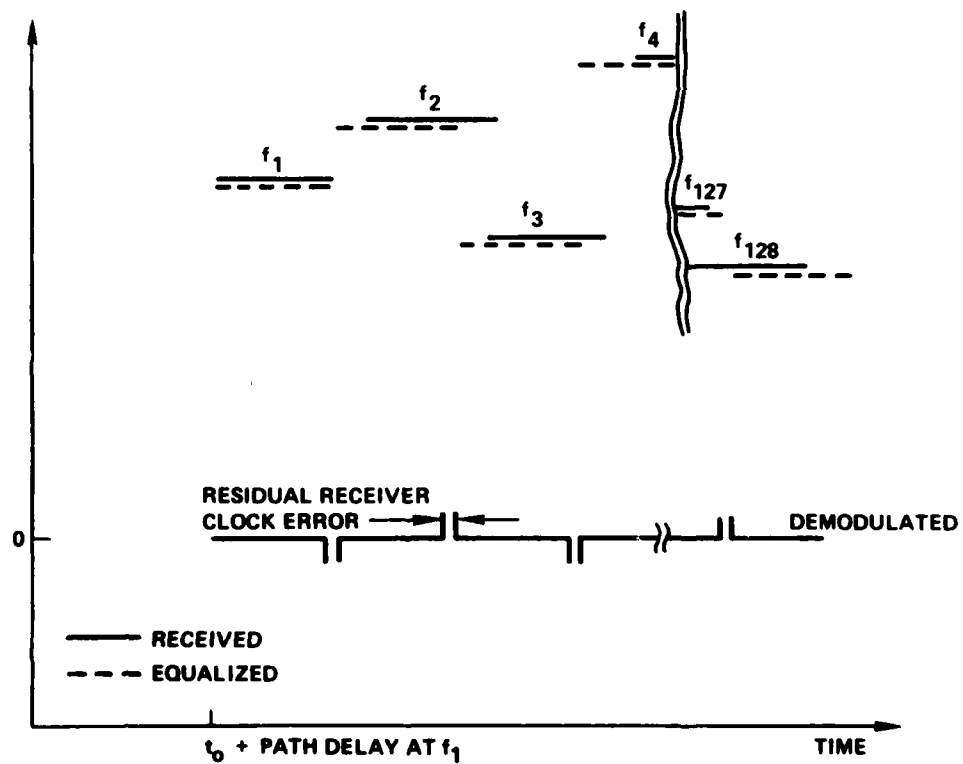
When frequency-hop spread spectrum is used for anti-jamming purposes, each dwell frequency is generally selected under the control of a pseudo-random process. For convenience, this system substitutes the probe scrambling format for random frequency selection. Scrambling of this type provides little protection against an adversary who wishes to predict where the next dwell will be. This is, however, not an issue in this exploratory program. Demonstrating the feasibility of scrambled FFH spread spectrum signalling will assure the success of pseudo-random FFH wideband HF communications.

As described in section 3.2, acquisition of the probe signal will result in a synchronization of the received signal and receiver clock to within a timing error commensurate with the unequalized channel dispersion. Figure 14 is a sketch of the time-frequency pattern of a single frame of the FFH waveform. The 128 frequencies arrive at the receiver at different times, depending on the dispersion of the path. This results in overlapping between dwells. The equalizer compensates for the dispersion and the overlap is removed. The FH pattern is then demodulated using a locally generated replica of the transmitted waveform. By refining the receiver clock timing to a small fraction of the dwell interval, a demodulated output is obtained which is phase-coherent and at a constant frequency throughout the frame, except for small intervals of transition due to the residual receiver clock error.

During these transition intervals, the demodulated frequency equals the frequency difference between adjacent dwells (from as little as 8 kHz to possibly as much as 1016 kHz). By integrating the demodulated output over the data bit interval, the effects of these short-duration frequency excursions are minimized.



14a: FFH Waveform Transmitted



14b: FFH Waveform Received, Equalized, and Demodulated

Figure 14: FFH WAVEFORM

### 3.6 Spectrum Analysis of Interference

In section 2.4, the impact of interference on the wideband HF user was discussed. The approach to the mitigation of interference is similar to that of dispersion and multipath: i.e. (1) measure the interference and then (2) compensate for it via the appropriate signal processing. A capability for interference measurement has been included in the test facility. These measurements will be used to develop a quantitative understanding of the nature of the interference which will later be used to formulate the appropriate signal processing.

The measurement of interference is achieved at both transmit and receive sites simultaneously. This will permit later comparative analysis and assessment of the efficacy of predistorting the transmitted signal in accordance with the measurements made at the transmit terminal. The measurement consists of an analog representation of the log of the magnitude of the spectrum within the band W.

The technique chosen is called panoramic analysis. It is shown in Figure 15. A local oscillator repetitively sweeps across the band W and is mixed with the receiver output. The predetection filter determines the analyzer resolution. This type of spectrum analysis was chosen because of its simplicity (as compared with, say, a filter bank or a Fourier transformer) and the fact that the LO waveform is easily obtained as an unscrambled version of the probe signal.

#### 3.6.1 Analyzer Resolution

By using the unscrambled probe signal as the LFM LO and by selecting the panoramic analysis technique, the following parameters apply:

- $W = 1024$  or  $512$  kHz
- $T_s = 1$  second

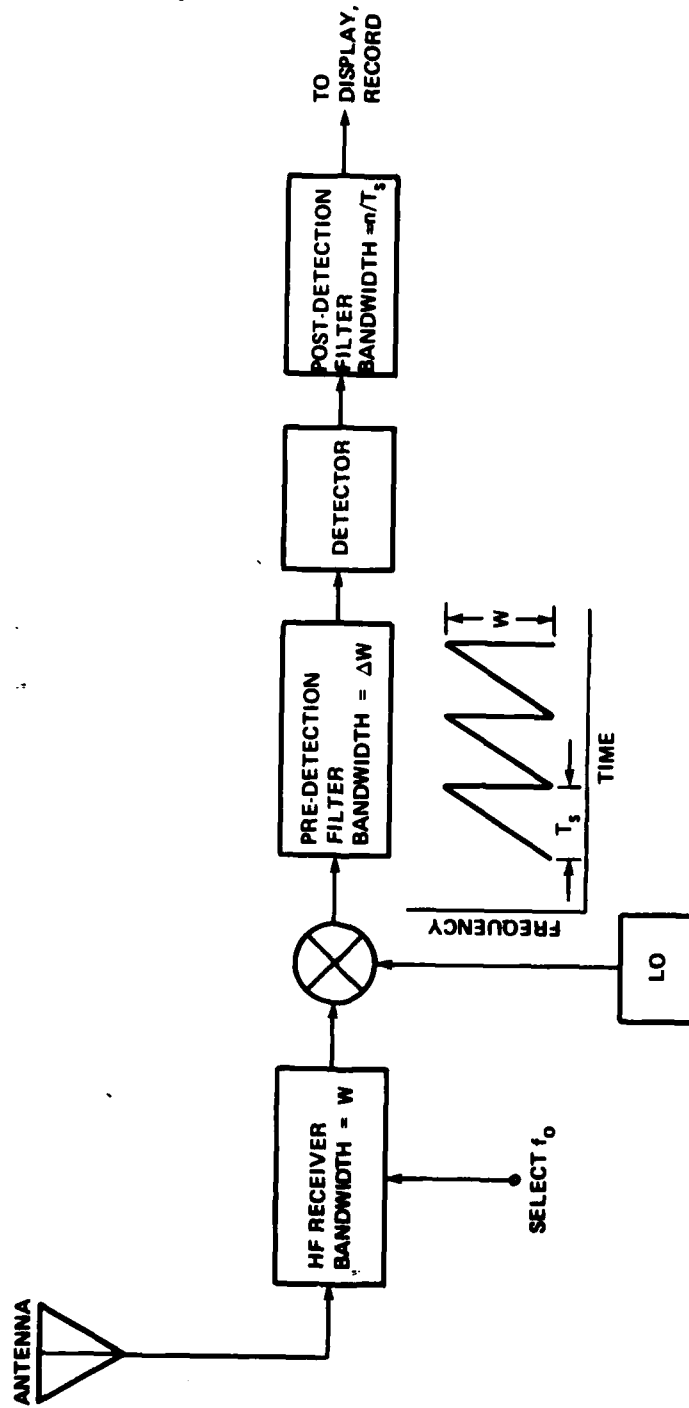


Figure 15: PANORAMIC SPECTRUM ANALYSIS



The time required for the LO to sweep across the resolution bandwidth  $\Delta W$  is

$$\Delta T_s = \frac{\Delta W}{W} T_s = \frac{T_s}{n} \quad (12)$$

where  $n$  is the number of resolution cells.  $\Delta T_s$  must not, however, be less than the build-up time of the predetection filter. This build-up time is at least equal to the reciprocal of  $\Delta W$  (and will be larger for high-order filters).

As a reasonable design value then

$$\Delta T_{s_{\min}} = \frac{1}{\Delta W} \quad (13)$$

combining (12) and (13)

$$\Delta W_{\min} = \sqrt{\frac{W}{T_s}}$$

which for our fastest sweep gives

$$\Delta W_{\min} = \sqrt{1024 \times 10^3} \approx 1 \text{ kHz}$$

and

$$n_{\max} \approx 1000$$

Because  $T_s = 1$  second, the postdetection bandwidth  $\frac{n}{T_s}$  is 1000 Hz. In our implementation (which is described in section 4.5) the predetection filter is a low pass filter, and its output is recorded on an audio cassette recorder. The postdetection analyzer output (after playback of the cassette) is recorded on a chart recorder or oscillograph. In order to reduce the bandwidth of the information to that which a reasonably priced chart recorder can reproduce while, at the same time, not exceeding the bandwidth of a portable cassette machine, the following parameters were chosen:

- $W = 6 \text{ kHz}$ , or twice the cutoff frequency of a 3 kHz audio cassette recorder,

- $n = \frac{1024}{6} = 170$ , and
- $\frac{n}{T_s} = 170 \text{ Hz}$ .

This resolution will resolve the majority of the narrowband interference carrier lines.

## SECTION IV

### TEST FACILITY DESCRIPTION

In this section the test facility is described in detail. Figure 16 is a simplified block diagram of the facility which shows the equipments for transmitting, receiving, equalizing, and recording wideband signals. The upper portion of Figure 16, the transmit terminal, is housed in the vehicle shown in Figure 17. The interior of the vehicle is shown in Figure 18. The left-hand photograph is of the program generator, synthesizer, exciter, and single-sideband (SSB) transceiver equipment while the right-hand photograph shows the transmitter amplifiers. A laboratory view of the receive terminal is shown in Figure 19. The left-hand rack contains the equalizer, synthesizer, and the program generator; the center rack contains the Hewlett-Packard (HP) minicomputer and digital displays, and the right-hand rack contains the receiver front and back ends. (The wideband tape recorder and the SSB transceiver are not shown in this view.)

Other equipment (not shown in Figure 16 but which will be described in this section) includes the timing circuits (including the terminal clock), time and frequency synchronization receivers, and the equipment for spectrum analysis of the interference.

#### 4.1 Equalizer

The circuitry used to implement the transversal filter equalizer (Figure 13) is described in this section. The parameters are as follows:

$$N = 128 \text{ or } 64$$

$$\tau = 1/W = \begin{cases} 0.9776 \text{ } \mu\text{s} & \text{for } W = 1024 \text{ kHz} \\ 1.953 \text{ } \mu\text{s} & \text{for } W = 512 \text{ kHz} \end{cases}$$

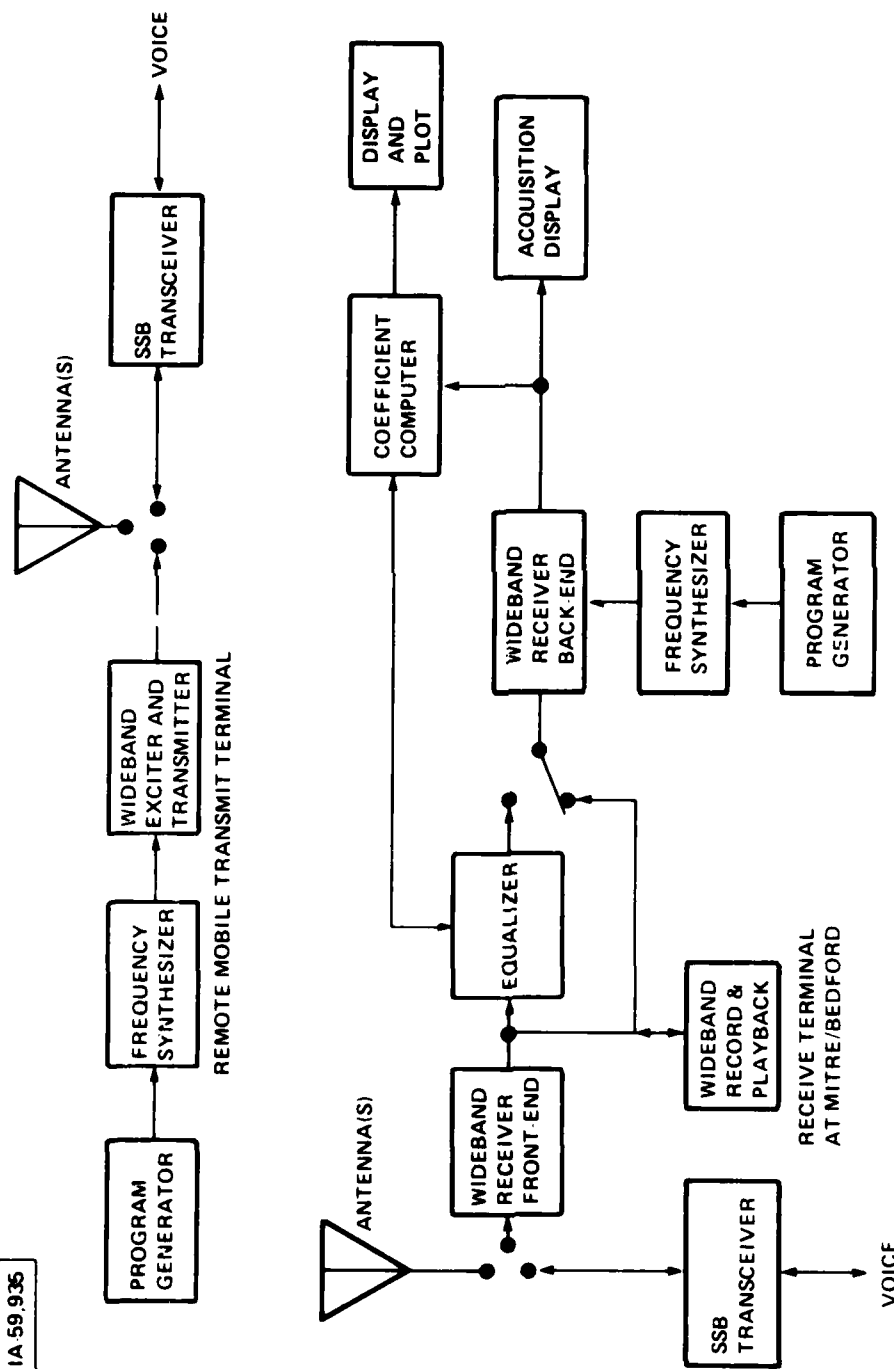


Figure 16: ONE-WAY WIDEBAND HF COMMUNICATIONS TEST FACILITY

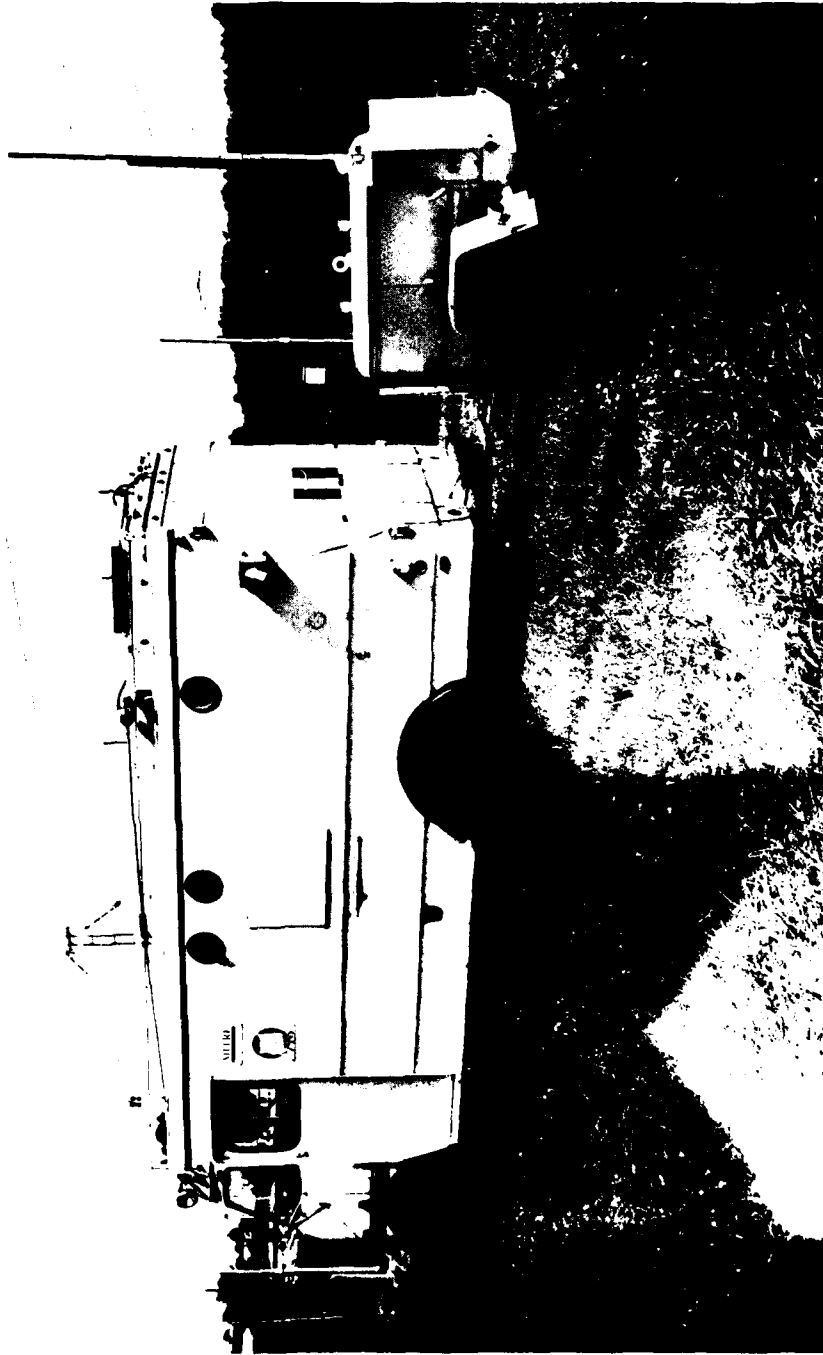


Figure 17: MOBILE TRANSMIT TERMINAL



Figure 18: TRANSMIT TERMINAL EQUIPMENT

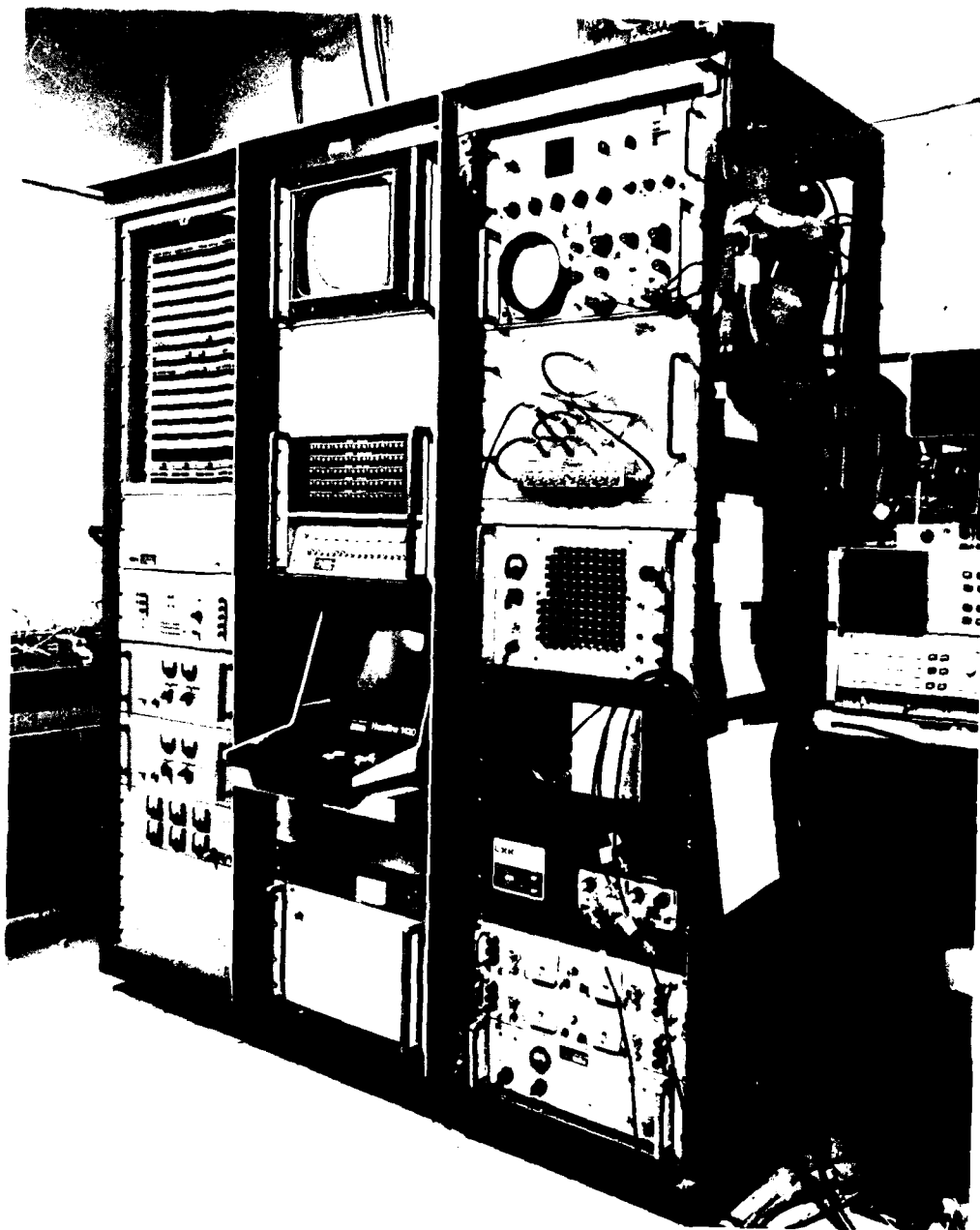


Figure 19: RECEIVE TERMINAL EQUIPMENT

$$T = N\tau = 62.5 \mu s, 125 \mu s, \text{ or } 250 \mu s$$

The equalizer is realized with all digital circuitry. It operates at baseband and interfaces with the receiver front end and back end via the equalizer interface unit. It also interfaces with the co-efficient computer and receives its clocking signal from the system timing generator.

#### 4.1.1 Equalizer Interface Unit

The equalizer interface unit is shown in Figure 20. In addition to interfacing with the equalizer, it connects with the VR-3700 wide-band tape recorder facility and provides two sets of LPF's. One set prevents aliasing in the equalizer input analog-to-digital converters (ADC) and the other eliminates harmonics from the equalizer output digital-to-analog converters (DAC). Both sets consist of four filters, two each for the in-phase (I) and quadrature (Q) channels, one for  $W = 1024 \text{ kHz}$ , and the other for  $W = 512 \text{ kHz}$ .

The input and output are both at an IF center frequency of 32.012 MHz and a crystal LO at this frequency is used for down-conversion to baseband and up-conversion back to the IF.

#### 4.1.2 Equalizer Detailed Description

The 128-stage tapped delay line transversal filter described in section 3.4 has been implemented digitally. The equalizer design is based on the use of recirculating memories and time-shared, high-speed digital multiplier/accumulators in a modular arrangement.

In the equalizer (shown in Figure 21) successive samples of  $\text{Re}[g(t)]$  and  $\text{Im}[g(t)]$ , in the I and Q input channels respectively, are weighted and summed to produce  $\text{Re}[h(t)]$  and  $\text{Im}[h(t)]$  in the I' and Q' output channels. Each output sample pair is the sum of 128 complex products



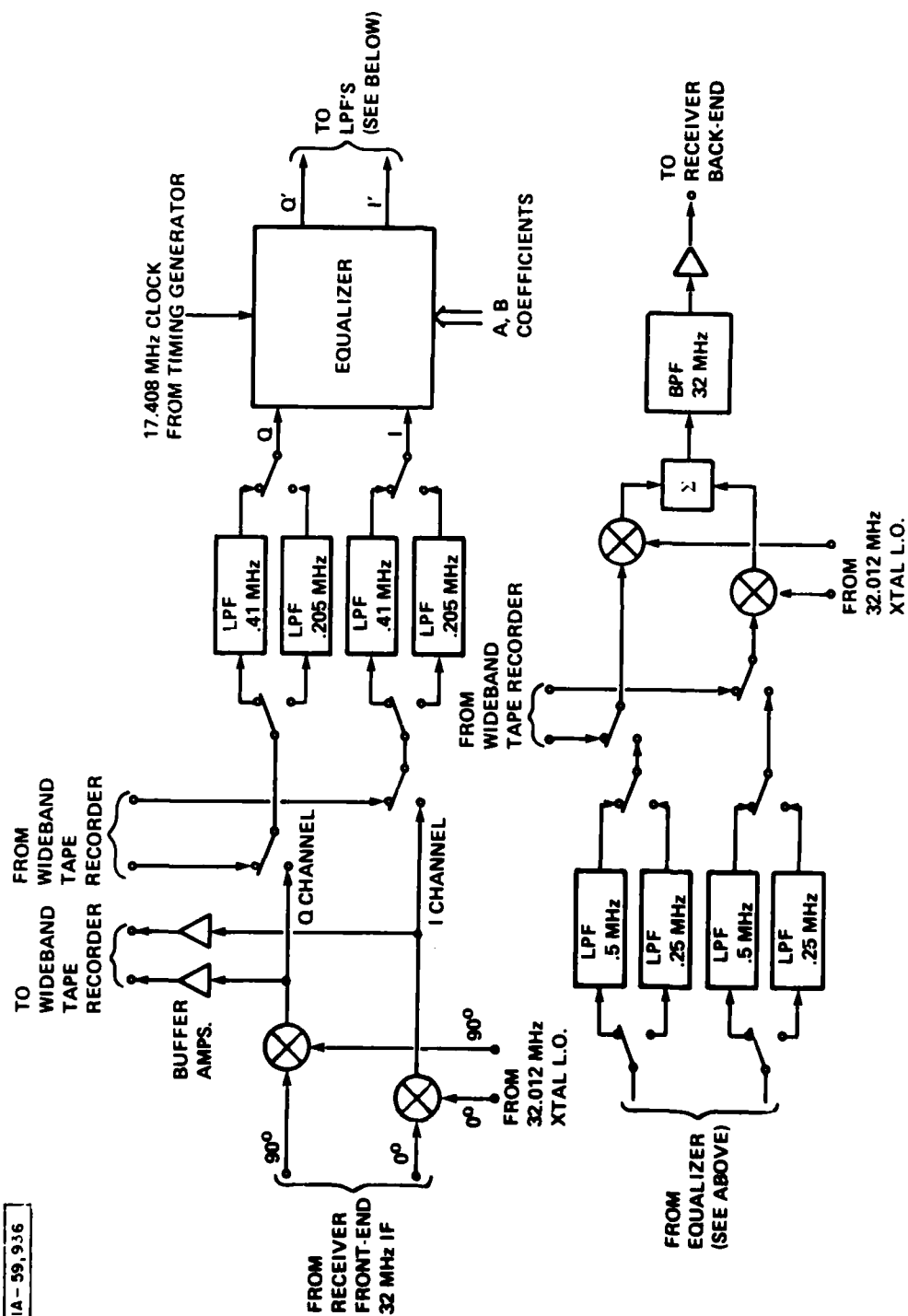


Figure 20: EQUALIZER INTERFACE UNIT

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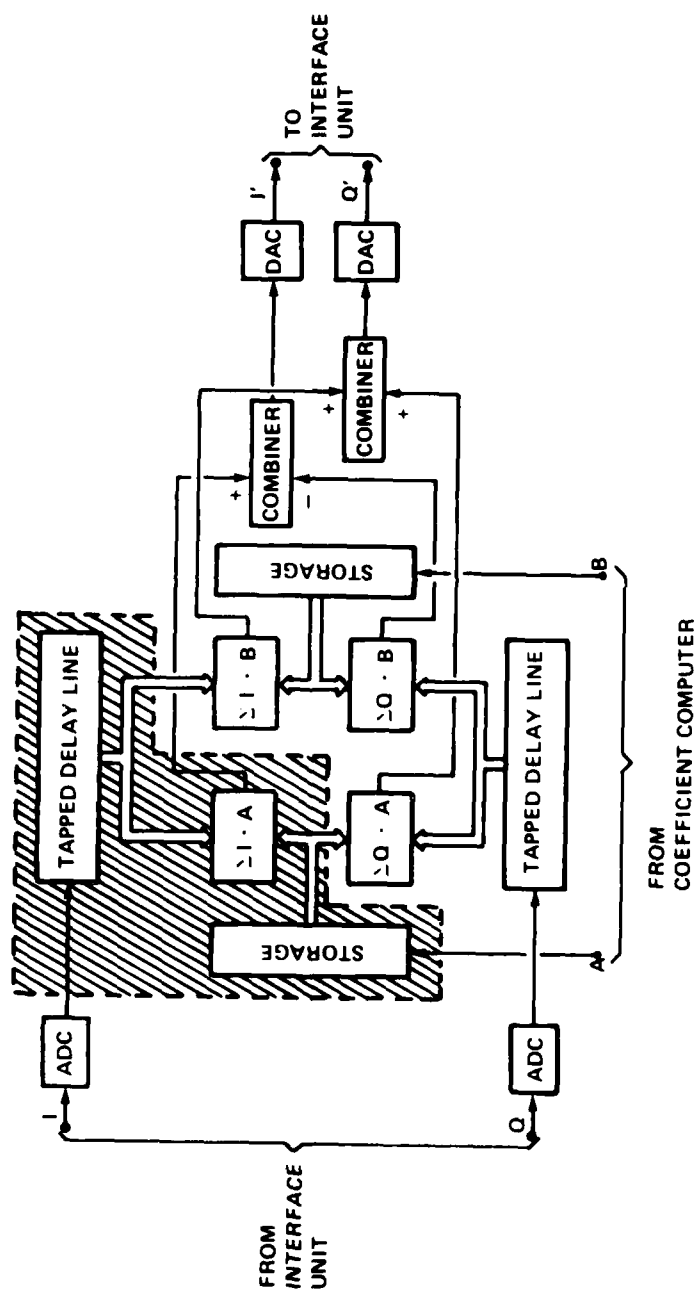


Figure 21: PROGRAMMABLE EQUALIZER

$$\begin{aligned}
I_t' + jQ_t' &= \sum_{k=1}^{128} (I_k + jQ_k) \cdot (A_k + jB_k) \\
&= \sum_{k=1}^{128} (I_k \cdot A_k - Q_k \cdot B_k) + j \sum_{k=1}^{128} (I_k \cdot B_k + Q_k \cdot A_k) \quad (14)
\end{aligned}$$

The four terms  $\sum I \cdot A$ ,  $\sum Q \cdot B$ ,  $\sum I \cdot B$ , and  $\sum Q \cdot A$  are computed first and then combined to give the two desired output sample values.

The I and Q channel input samples are quantized to 8 bits in amplitude. The sampling rate is either 1024 or 512 kHz depending on the choice of W. The  $A_k$  and  $B_k$  coefficients are also 8-bit words which are entered into the equalizer at a rate determined by the coefficient computer but synchronized with the equalizer sampling clock.

Figure 22 shows the shaded area of Figure 21 in detail. It includes the I-channel digital tapped delay line and the A-channel coefficient storage register (which are actually identical in their implementation). Eight multiplier/accumulators each compute the sum of 16 products by a time-sharing arrangement involving the local recirculation of 16 stages of memory content. The 16-bit products from the multipliers are rounded to 12 bits. The accumulators send 16 bits to the 8-to-1 MUX. The final accumulator combines the 8 MUX outputs and sends 19-bit data to the combiner. Before entering the DAC's (see Figure 21), a front panel controlled scaler circuit selects the best 10 bits of the 20-bit combiner output.

A pipeline structure is used in the implementation of the equalizer to achieve the high throughput speed. A 1-bit or 2-bit pipeline delay permits processing the preceding result while simultaneously acquiring the next input sample. In the digital multiplier, the 8-bit multiplier and 8-bit multiplicand are loaded into their "x" and

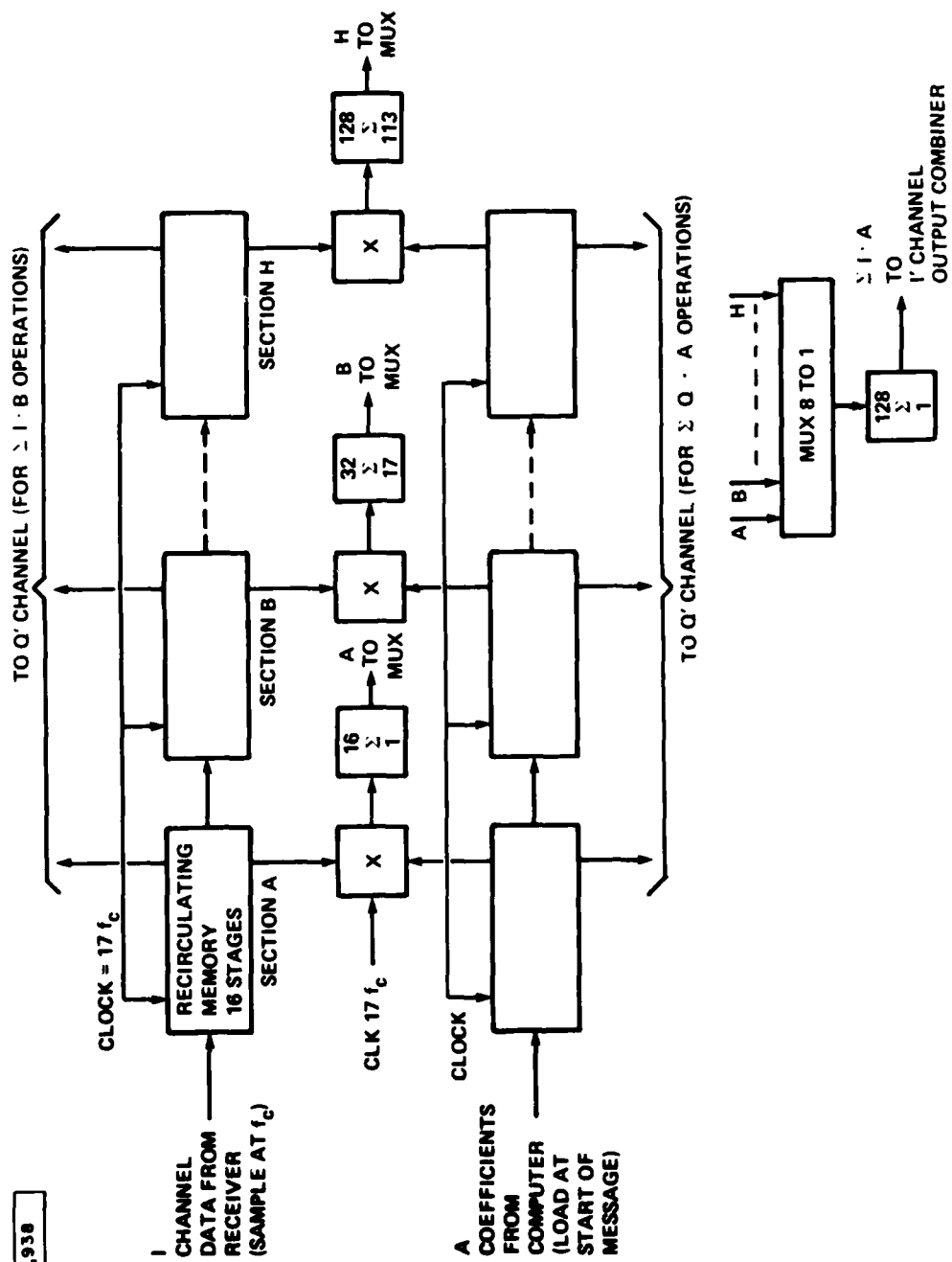


Figure 22: EQUALIZER DETAILS (SHADED REGION OF FIGURE 21)

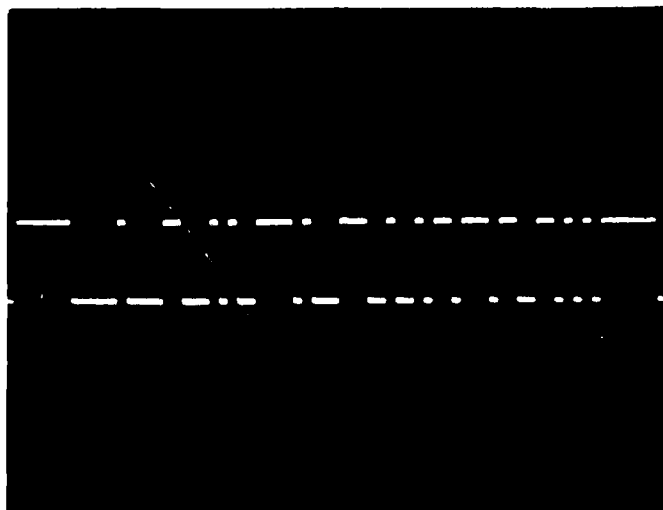
"y" registers while the product of the previous operands is simultaneously loaded into its output register. Because of the delay in the output register, latches or storage registers are required between the multiplier and accumulator to achieve the high throughput speed.

#### 4.1.3 High-Speed Digital Multiplier Problem

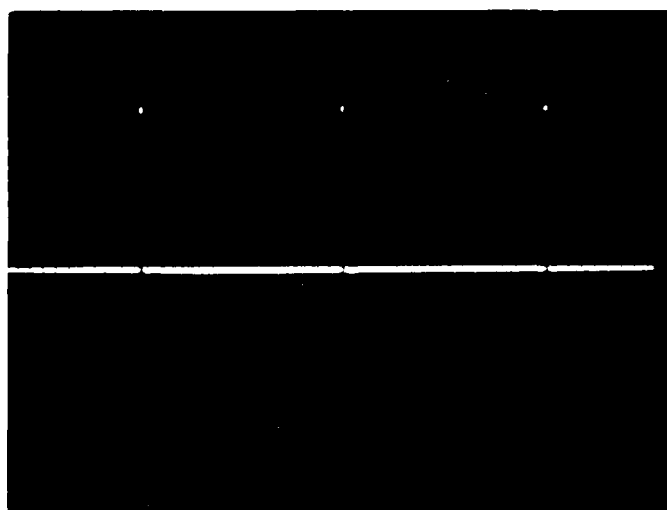
The time-sharing multiplication scheme described above requires that each 8-bit by 8-bit multiplication be completed within 57.4 nanoseconds which is one seventeenth of the ADC sampling interval. TRW model MPY-8HJ-1 devices are employed which were specified to have a typical 25°C multiply time of 45 ns and a maximum multiply time over 0°C to 60°C of 60 ns. We expected that the maximum multiply time at room temperature would be less than the required 57.4 ns, but we discovered that under certain data-dependent conditions this was not the case. A cooperative effort with TRW uncovered the cause and led to the resolution of the difficulty. TRW subsequently supplied us with a new set of upgraded devices which meet our requirements. We have also developed a circuit clocking modification which permits slower operation (by a few nanoseconds) and provides additional protection against multiplier speed variations with age, temperature and power supply changes.

#### 4.1.4 Equalizer Testing

The equalizer has been extensively tested to demonstrate its capabilities. One example of this testing is to use one tapped delay line (say the I channel) and one storage register (say the A storage register) to compute a correlation function ( $\sum_{k=1}^{128} I_k \cdot A_k$ ). Special built-in test equipment is used to load a maximum-length sequence binary phase code for  $m = 63$  into the first 63 stages of the storage register (the last 65 stages are set to zero). The same  $m = 63$  code is then repetitively clocked into the tapped delay line at 1024 kHz. Figure 23 shows the results of this test. Figure 23a displays slightly



23a:  $m = 63$  Maximum Length Sequence



23b: Autocorrelation Function for  $m = 63$  Sequence as Seen at  
I' Output  $W = 1024$  kHz

Figure 23: EQUALIZER TEST RESULTS

more than one cycle of the repeating sequence, and Figure 23b shows the circular correlation function seen at the equalizer's I' channel output when the B register and/or Q channel are set to zero. Relative to the zero voltage level, the mainlobe is +63 and all sidelobes are -1 as expected.

#### 4.2 Coefficient Computer

The coefficient computer receives frequency-domain measurements of the transmitter-to-receiver path, computes the appropriate frequency-domain equalizer coefficients, and transforms these coefficients into the time domain as complex tap weights for use by the system's transversal filter equalizer. A block diagram showing the coefficient computer and its various interfaces is shown in Figure 24.

The ADC's sample the received one-second probe frame at a higher rate than the required 128 Hz. This oversampling avoids the excessive circuit delays which would be associated with anti-aliasing filters at a 128 Hz sampling rate. The redundant samples are combined to yield the required 128 complex samples of the channel transfer function. These are then unscrambled using information from the synthesizer programming circuits before the coefficients are computed. The complex tap weights for the equalizer are available approximately 200 ms after the end of the one second probe ( $T_p$ ) is received. Outputs for both 128- and 64-point equalization are provided.

##### 4.2.1 Coefficient Computer Software

As shown in Figure 25, the coefficient computer software performs the following functions:

- The ADC samples are combined and unscrambled thus providing a sampled data, rectangular coordinate representation ( $C_\ell$ ) of the channel transfer function  $\{C(\omega)\}$ .
- The frequency domain weights ( $Z_\ell = X_\ell + jY_\ell$ ) are computed according to equation (6).

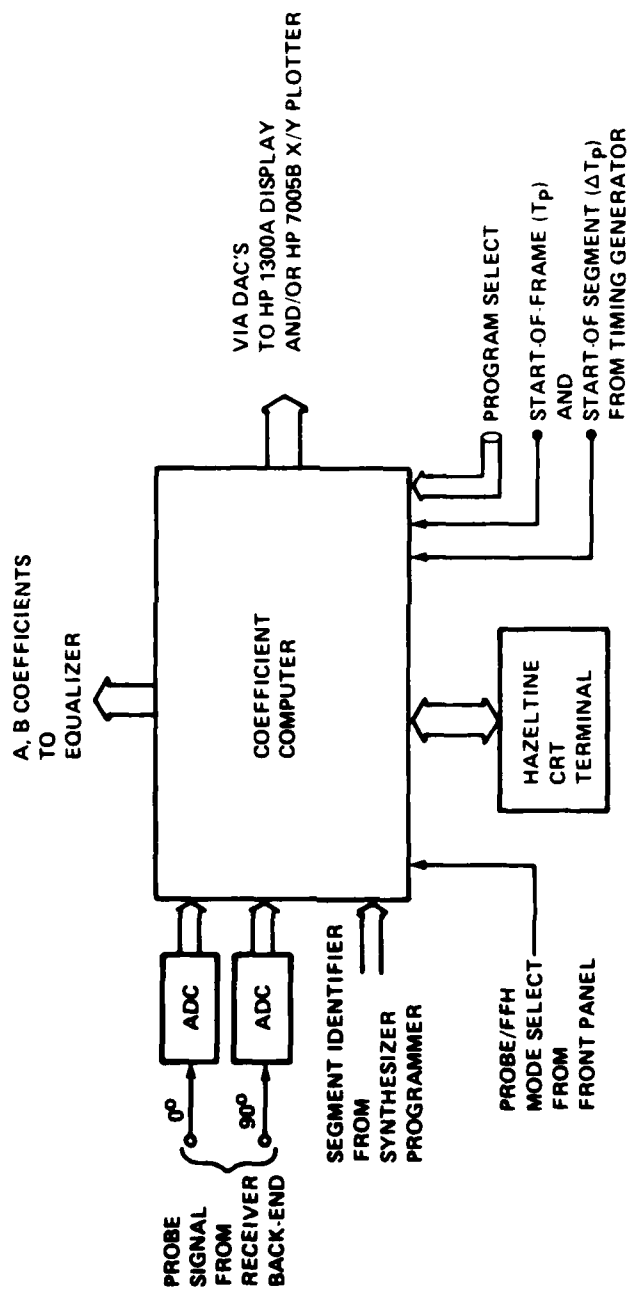


Figure 24: COEFFICIENT COMPUTER INTERFACES



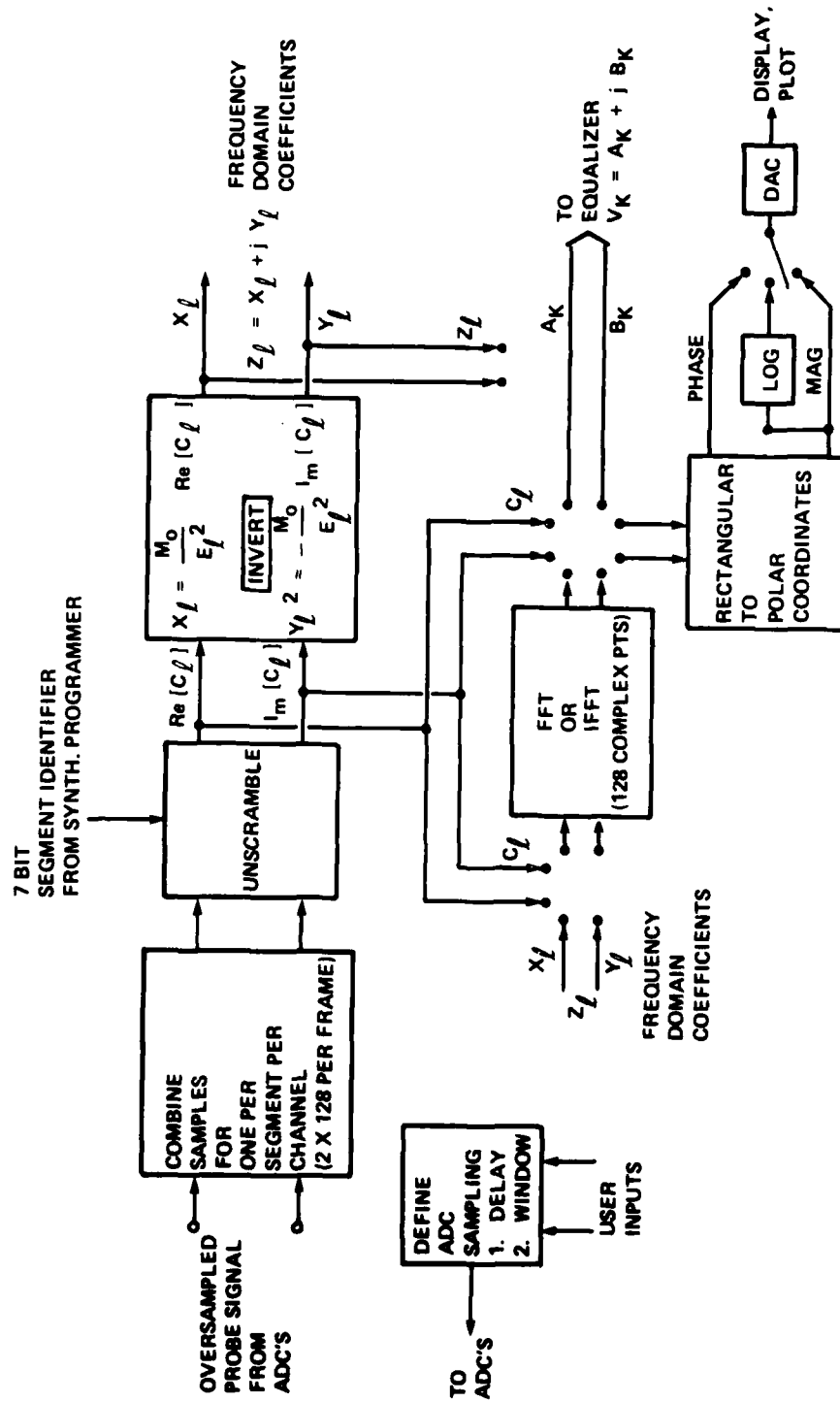


Figure 25: COEFFICIENT COMPUTER FUNCTIONAL DESCRIPTION

- The equalizer coefficients ( $V_k = A_k + jB_k$ ) are computed. They are the inverse FFT of the  $Z_{\ell}^{k_{\text{set}}}$ .
- The coefficients are loaded into the equalizer A and B storage registers.
- The FFT section can also be used to provide the transform of  $C_{\ell}$  which is the channel impulse response.
- For display and/or plotting, rectangular coordinates are mapped to polar coordinates, thus providing the magnitude and phase of  $C_{\ell}$ ,  $Z_{\ell}$ ,  $V_k$ , or the transform of  $C_{\ell}$ . The log of the magnitude is also provided.

#### 4.2.2 Coefficient Computer Hardware

All software functions are presently implemented on an HP 2115A minicomputer. This equipment is in the process of being upgraded to the newer HP 21MX. An HP 1300A large-screen CRT display, an HP 7005B X/Y plotter, and a Hazeltine 1420 CRT terminal constitute the associated peripherals.

### 4.3 Time/Frequency Circuits

In this section the programmed frequency synthesizers and the equipment and circuits used to supply all the necessary timing signals and local oscillator frequencies is described. The block diagram of the time/frequency circuits is shown in Figure 26. (The frequency standard, clock generation, and timing generator circuitry is not shown in the test facility system block diagram of Figure 16.)

#### 4.3.1 Frequency Standards

The frequency and timing generators at each terminal of the test facility are all referenced to a standard one-megahertz signal. This standard is the one-megahertz quartz crystal oscillator within the Rockland synthesizer. The HP synthesizer is used to provide a band-set local oscillator function. Its driver also is used for local

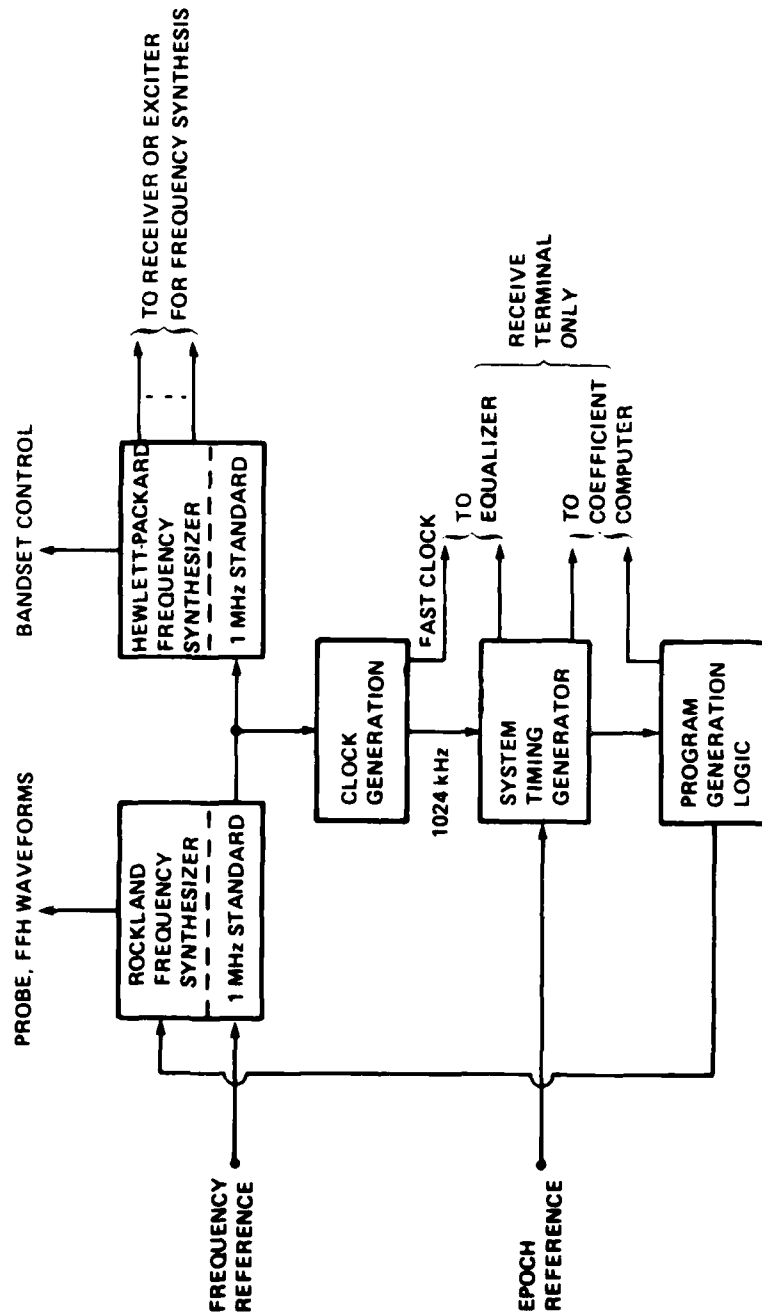


Figure 26: TIME/FREQUENCY CIRCUITS

oscillator frequency synthesis. Interterminal frequency synchronization will eventually be achieved by referencing each terminal's standard to a common external reference signal as provided by standard frequency broadcasts from WWVB. In the present stand-alone configuration, the standard oscillator's long-term stability of one part in  $10^9$  per day results in a need to occasionally readjust one of the standards.

#### 4.3.2 Clock Generation

The clock rates employed in the system are derived using phase-locked-loop (PLL) techniques. The basic system clock rate is equal to the higher value of W of 1024 kHz while the system "fast clock", used in the equalizer, is seventeen times this value, i.e., 17.408 MHz. A functional block diagram of the circuitry involved is shown in Figure 27. Epoch synchronization between the two terminals is achieved using the procedure described in section 4.6 and implemented by interrupting the master clock for a selected number of clock pulses.

#### 4.3.3 System Timing Generator

The digital logic used to generate the required timing waveforms is shown in Figure 28. The 1024 kHz master clock is divided down in two chains; the following outputs are provided:

- probe and LFM or FFH mode start-of-frame ( $T_p/T_f$ ).
- probe and LFM start-of-segment ( $\Delta T_p$ ) or FFH start-of-dwell ( $\Delta T_f$ ) ( $2\Delta T_f$  is also used to increment the  $\Delta f$  counter in the probe and LFM modes.),
- load command to the Rockland synthesizer,
- reset commands to the  $\Delta f$  counter, and
- one Hertz "ticks" for time marks on recordings and for WWV/CHU synchronization.

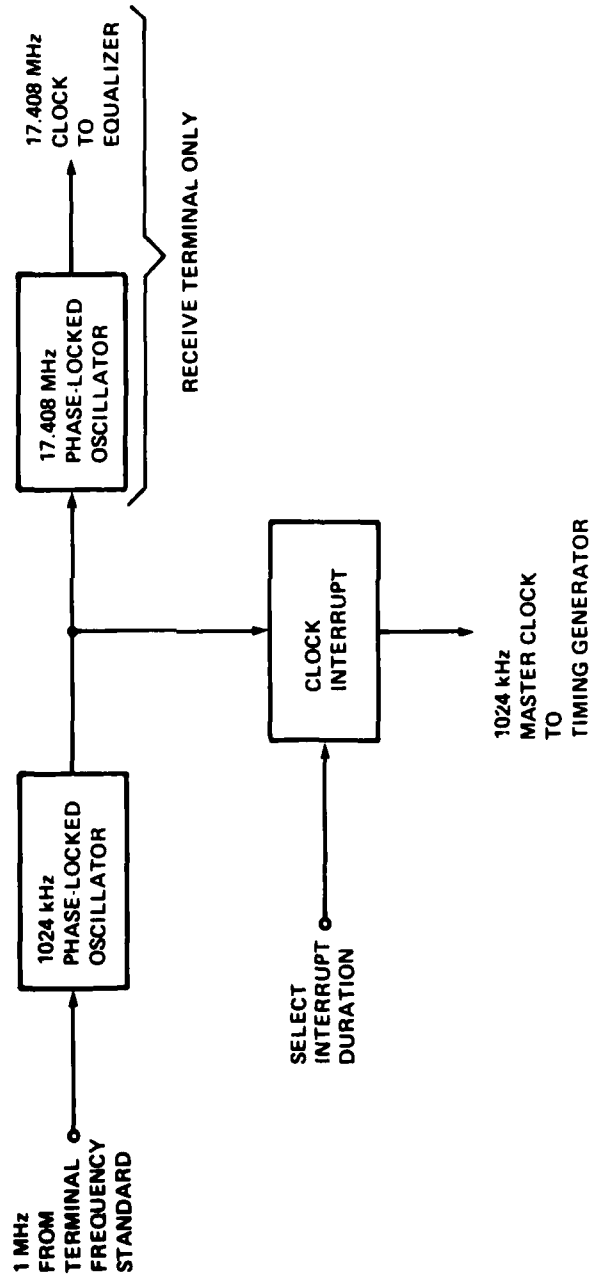
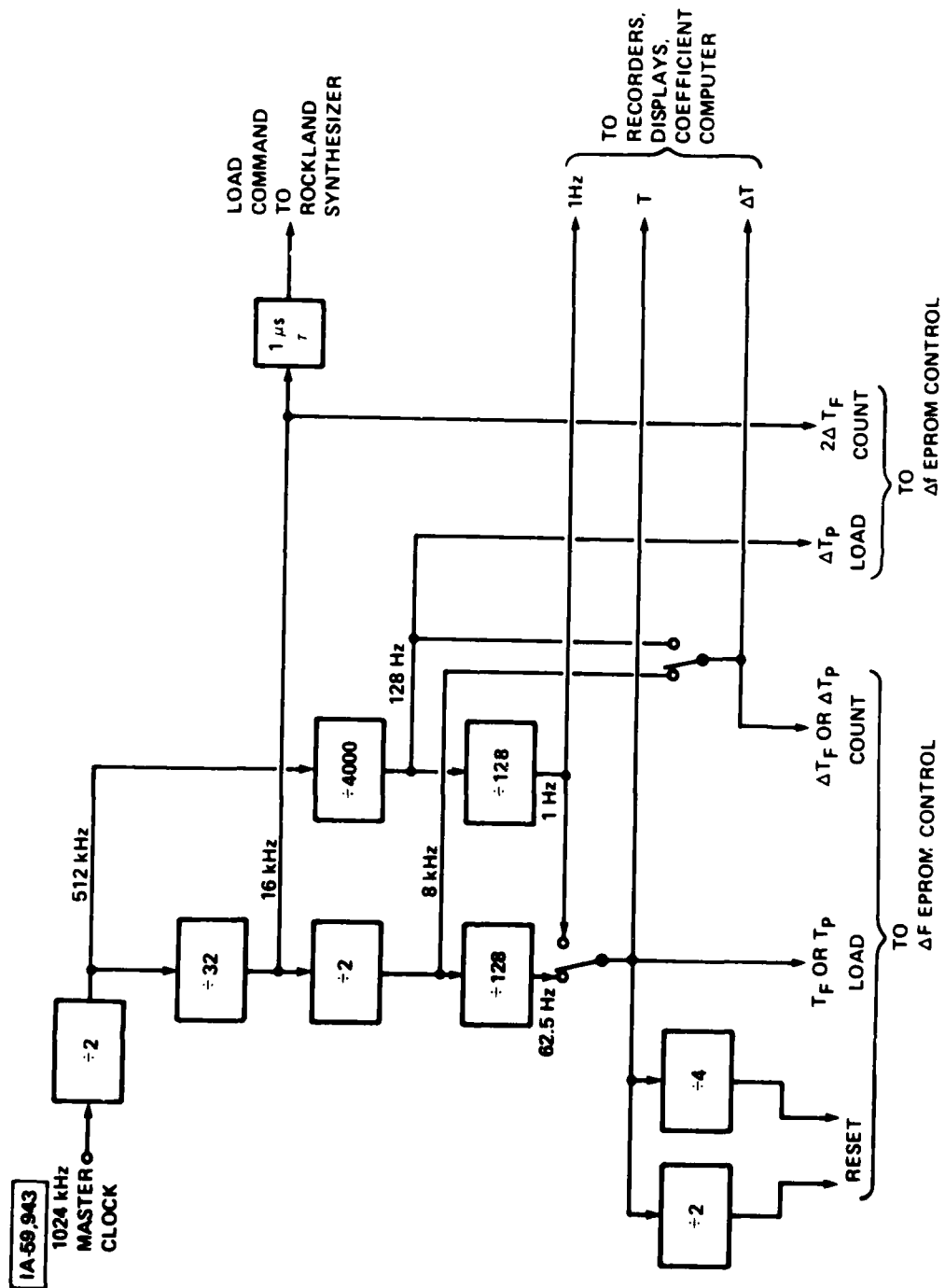


Figure 27: CLOCK GENERATION CIRCUITS



One of two timing regimes can be selected by a front-panel switch. A slow timing regime based on a one second frame time ( $T_p$ ) provides  $\Delta T_p = \frac{1}{128}$  second and  $\Delta t = 125 \mu s$ . In the fast timing regime, the frame time ( $T_f$ ) is 16 ms and  $\Delta T_f = 125 \mu s$ .

#### 4.3.4 Program Generation

The Rockland synthesizer employs a 32-bit digital word to select the desired output frequency. The method of control used permits the 13 least significant bits to be set to zero while the remaining 19 bits are programmed. Because of the particular format used by Rockland for synthesizer control, deriving the control programs is done by pre-programming electrically programmable read only memories (EPROM's) with the EPROM contents being read out and fed to the Rockland control inputs under the command of conventional binary logic.

A functional block diagram showing the program generation logic appears in Figure 29. The  $\Delta F$  EPROM contains eight programs, each containing 128 8-bit words, for coarse frequency ( $\Delta F$ ) selection. The  $\Delta F$  values are either 4, 8 or 16 kHz apart, depending on the selected values of bandwidth (W) and number (N) of segments/dwells. The  $\Delta F$  EPROM address counter is clocked by  $\Delta T_p$  in the probe and LFM modes and by  $\Delta T_f$  in the FFH mode.

The  $\Delta f$  EPROM contains three programs, each containing 64 11-bit words, for fine frequency positioning. The  $\Delta f$  values are 64 Hz, 128 Hz, or 256 Hz apart. All three programs cause the synthesizer to generate a staircase approximation to LFM within a probe or LFM mode segment. The  $\Delta f$  address counter changes  $\Delta f$  every 125  $\mu s$  ( $\Delta t$ ). The phase error relative to true LFM is given by

$$\Delta \theta = \pm \frac{1}{8} \Delta t \Delta f \text{ in revolutions} \quad (15)$$

For  $\Delta t = 125 \mu s$  and  $\Delta f = 256 \text{ Hz}$ ,  $\Delta \theta = \pm 1.44$  degrees. In the FFH mode,  $\Delta f$  is held fixed at a value corresponding to approximately  $\frac{\Delta F}{2}$ .

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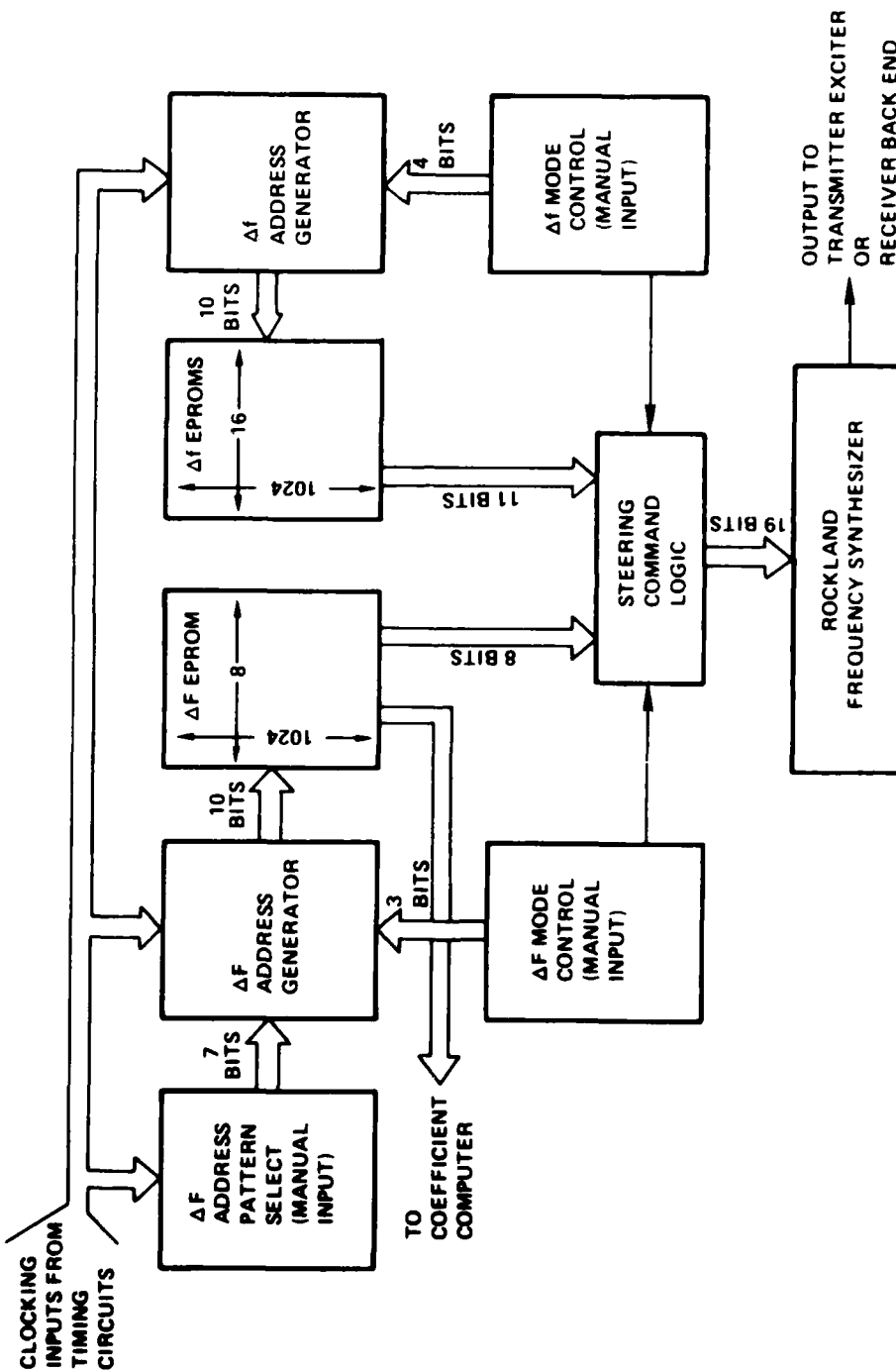


Figure 29: SYNTHESIZER PROGRAM GENERATION LOGIC



An eight-position front-panel rotary switch for program selection together with the slow-fast timing regime switch is used to select the desired mode (probe, LFM, or FFH), bandwidth (W), and number of segment/dwells (N). Each rotary switch position corresponds to a program in the  $\Delta F$  EPROM. Table I lists the synthesizer output waveform parameters for the various control settings.

The FFH waveforms available in program G provide a means for moderate band signalling which can be used under favorable ionospheric conditions without the need for channel equalization. Program H is essentially the same as program A except the frequency dwells on the edges of the band have been translated to the central 768 kHz so as to provide a means of FFH signalling which does not exceed the down-converter LPF cutoff frequency of  $\pm 400$  kHz.

#### 4.3.5 Frequency Synthesizers

The system employs two Rockland model 5100 programmable frequency synthesizers - one in the transmit terminal for probe, FFH, and LFM waveform generation, and one in the receive terminal for probe and FFH frequency demodulation as well as for LFM waveform generation. The synthesizers feature direct digital synthesis, and were selected because of their full programmability and fast switching capability over a greater-than-2-MHz band.

#### 4.4 Transmitter and Receiver Analog Circuits

In this section the antennas, transmit terminal exciter and power amplifier chain, and the receive terminal front end and back end are described. Since the exciter and the receiver front end are duals of one another and use identical circuit modules, they are discussed first.

TABLE I. SYNTHESIZER FREQUENCY CONTROL PROGRAMS

Program Select	Timing Regime	Mode	$\Delta F$ EPRM	$\Delta F$ EPROM	$\Delta F$ EPROM	W (kHz)	N	$\Delta F$ (kHz)	$\Delta f$ (Hz)	Comments
A	SLOW	PROBE	Scrambled	"	LPM	1024	128	8	128	
	FAST	FFH								
B	SLOW	PROBE	"	"	LPM	512	64	8	128	
	FAST	FFH								
C	SLOW	PROBE	"	"	LPM	1024	64	16	256	
	FAST	FFH								
D	SLOW	PROBE	"	"	LPM	512	128	4	64	
	FAST	FFH								
E	SLOW	LPM	LPM	LPM	LPM	1024	128	8	128	For Spectrum Analysis of Interference
	FAST	-								
F	SLOW	LPM	LPM	LPM	LPM	512	128	4	64	As above
	FAST	-								
	SLOW	-	Scrambled	Scrambled	LPM	1024	See Comment	8	-	16 sets of 8 scrambled dwells each covering 64 kHz
	FAST	Special FH								
H	SLOW	-	Scrambled	Scrambled	LPM	768	128	8	-	Outside dwells folded in for W = 768 kHz
	FAST	Folded FFH								

#### 4.4.1 Exciter and Receiver Front End

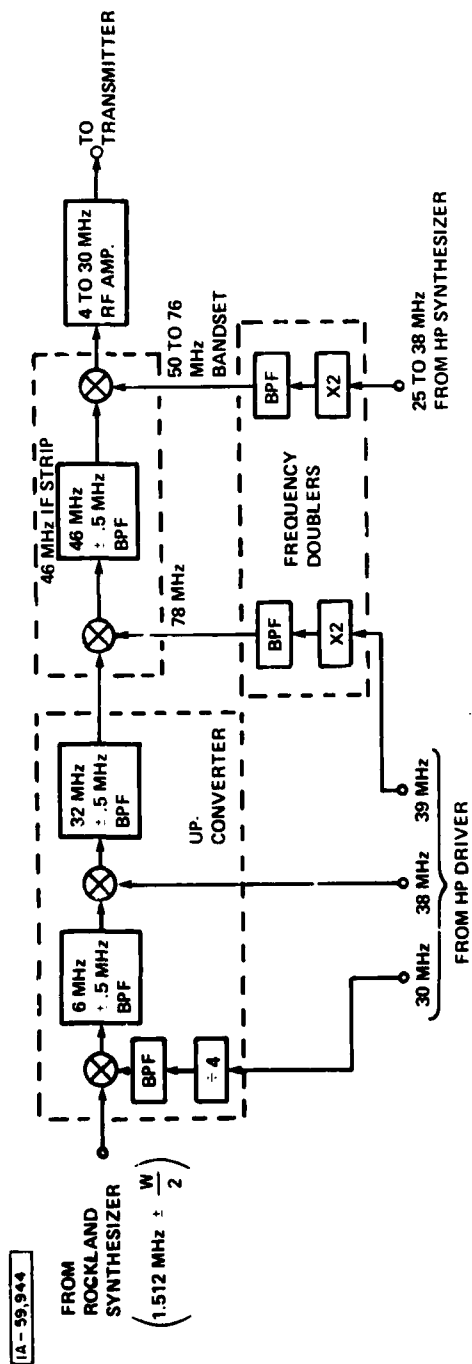
Because of the wideband signals used in the test facility, no conventional equipments or IF and other circuit modules that met the requirements of the exciter and the receiver front end were available. In the exciter it is necessary to take the up to 1024 kHz bandwidth waveform centered at 1512 kHz from the Rockland synthesizer and to frequency-translate this signal to any section of the HF band between 4 and 30 MHz. This is done first by up-converting to a 32 MHz IF center frequency and then by double-converting via a 46 MHz IF to the desired HF band center frequency.\*

In the receiver, the opposite is required. The signal from anywhere in the 4- to 30-MHz HF band is double-converted via a 46-MHz IF strip to an IF at 32 MHz. Here the signal is distributed to (1) the equalizer and the wideband tape recorder facility via the equalizer interface unit, (2) the receiver back end for down-conversion and demodulation of the probe signal, and (3) the spectrum analysis circuitry (see section 4.5). The block diagrams of the exciter and the receiver front end are shown in Figure 30. Because of the more severe dynamic range requirements in the receiver, out-of-band noise and interference are removed at the antenna output with a tunable preselector band pass filter.

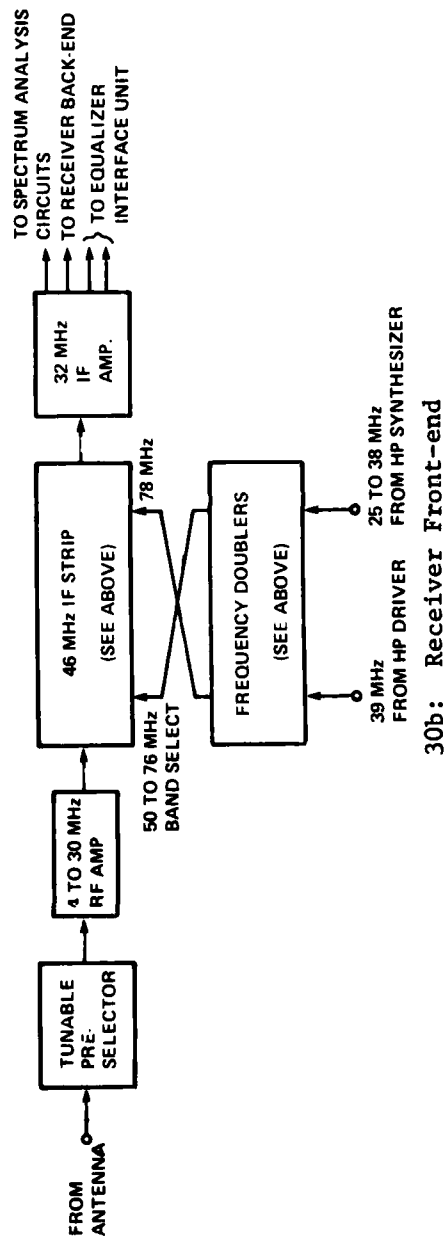
In both terminals an HP 5110A frequency synthesizer and its associated driver are used for local oscillator generation (see also Figure 26). The up-converter is also used in the receive terminal spectrum analysis circuits. All interunit connections are via a patch panel which is used in the transmit terminal to reconfigure the exciter circuit modules into a receiver for spectrum analysis.

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\*Here and elsewhere in this report the exact center frequencies of 32.012 MHz and 45.988 MHz are approximated for readability.



30a: Exciter



30b: Receiver Front-end

Figure 30: EXCITER AND RECEIVER FRONT-END

The receiver front end has a measured dynamic range of 60 dB (from 1 dB compression to the noise level). Worst-case spurious signals are 42 dB down when the maximum input signal is applied. The exciter output spectrum for  $W = 1024$  kHz and an RF frequency of 6 MHz is shown in Figure 31. The third harmonic is 45 dB down. For higher RF frequencies the harmonic level decreases. For lower frequencies it increases to 30 dB down at 4 MHz.

#### 4.4.2 Transmitter Amplifier Chain

The transmitter chain can deliver at least 500 watts of RF output power anywhere in the 4- to 30-MHz frequency range. It consists of three separate units: a 4-watt broadband low-power driver, a 100-watt intermediate power broadband driver, and a 500-watt high-power broadband amplifier.

All amplifiers operate in a linear (class A or B) mode and use wideband circuitry. Therefore, no adjustments are required when changing from one frequency to another. For RF center frequencies below 10 MHz, the third harmonic is within the 4- to 30-MHz band. Two high-level low pass filters are therefore used in this frequency region to reduce the harmonic content. The block diagram of the transmitter chain is shown in Figure 32.

The measured output spectrum at 500 watts and a RF center frequency of 6.6 MHz is shown in Figure 33.

#### 4.4.3 Antennas

The following antennas are available for use with the test facility:

##### A. Receive Terminal Antennas

- Dual 16-element phased array. Boresites  $45^{\circ}$ ,  $255^{\circ}$ . Steerable over  $\pm 45^{\circ}$ .

1A-59,945

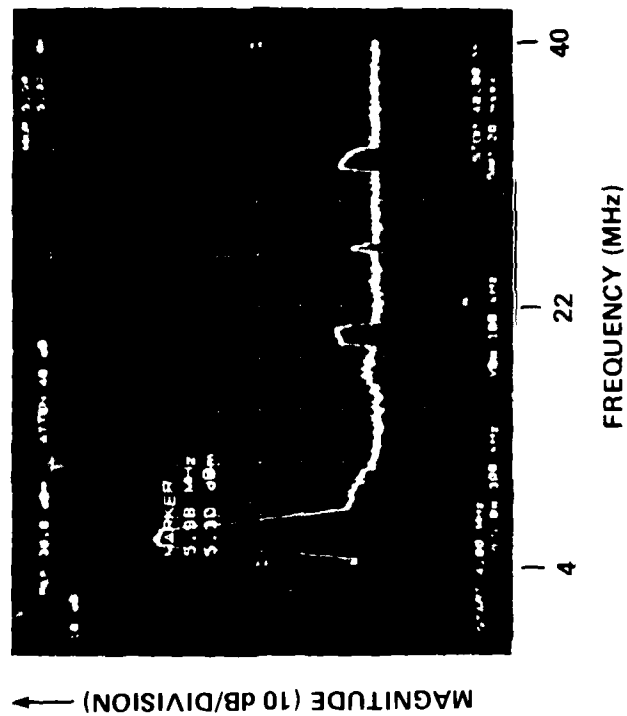


Figure 31: EXCITER OUTPUT SPECTRUM; CENTER FREQUENCY = 6.0 MHz

1A-59,946

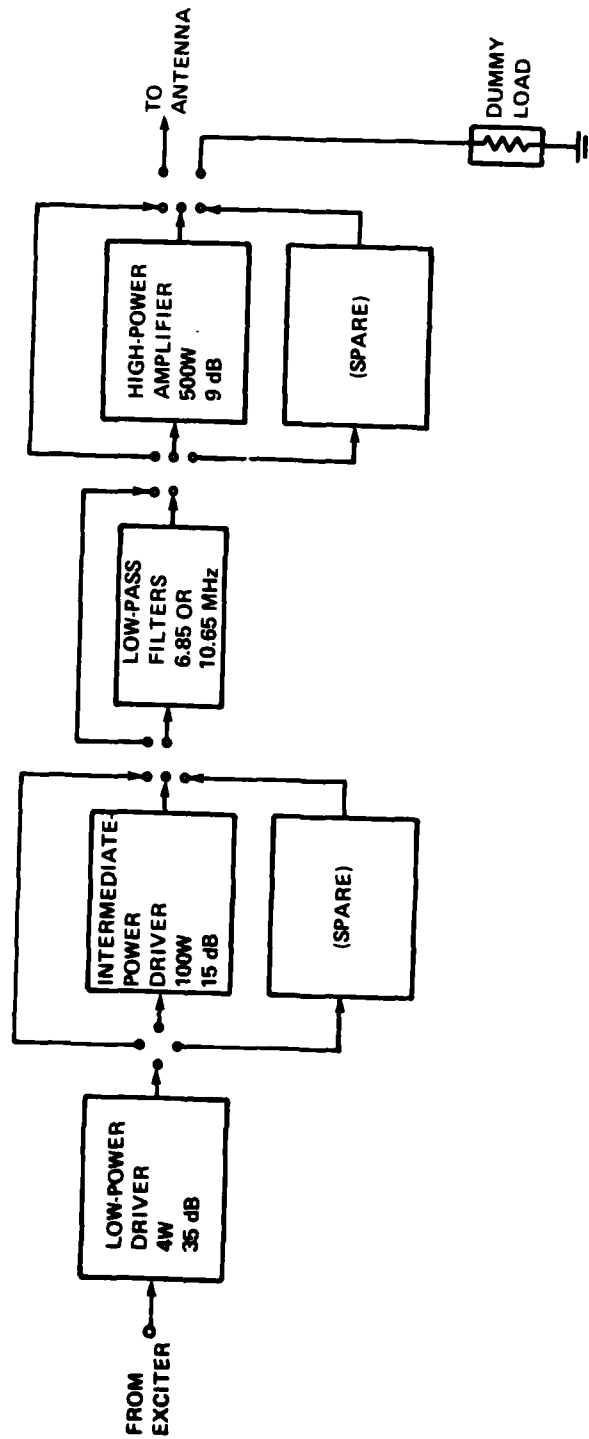


Figure 32: TRANSMITTER AMPLIFIER CHAIN

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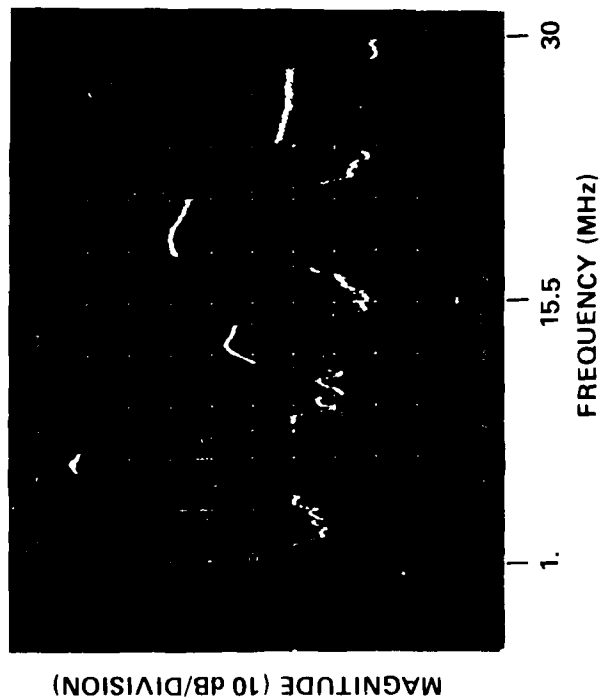


Figure 33: TRANSMITTER OUTPUT SPECTRUM: 500 WATTS, CENTER FREQUENCY 6.6 MHz



- Oblique log-periodic antenna (LPA). Boresite  $200^{\circ}$ . 6.5 to 30 MHz. Vertical polarization.
- Oblique LPA. Boresite  $200^{\circ}$ . 6.5 to 30 MHz. Horizontal polarization (used with SSB transceiver).
- Whip. (Used with WWV/CHU receiver).
- Zenith-directed LPA. 4 to 13 MHz. (Used for reception of Boston-Hill transmissions).

#### B. Transmit Terminal Antennas

- Transportable, oblique LPA. 12- to 22-MHz or 16- to 30-MHz Horizontal polarization.
- Bowtie. Located at Boston Hill. 1-MHz bandwidth at 5.5 MHz.
- Bowtie. Located at Boston Hill. 1-MHz bandwidth at 6.6 MHz.
- Bowtie. Located at Boston Hill. 1-MHz bandwidth at 7.7 MHz.
- Whip for SSB Transceiver
- Whip for WWV/CHU Receiver

The bowtie antennas were designed and built under this project. They are essentially broadband dipoles, horizontally polarized so as to give a modest zenith directivity.

#### 4.4.4 Receiver Back End and Display

This unit is shown in Figure 34. It accepts a 32-MHz IF wide-band signal from either the receiver front end or the qualizer interface unit. In-phase and quadrature IF signals are generated at 1.512 MHz by down-conversion with a 30.5-MHz LO. The probe or FFH frequency modulation is then removed in a second down-converter to baseband which uses the Rockland synthesizer as its local oscillator. At baseband, 100 kHz LPF's eliminate the IF and LO signals. The output

1A-55,948

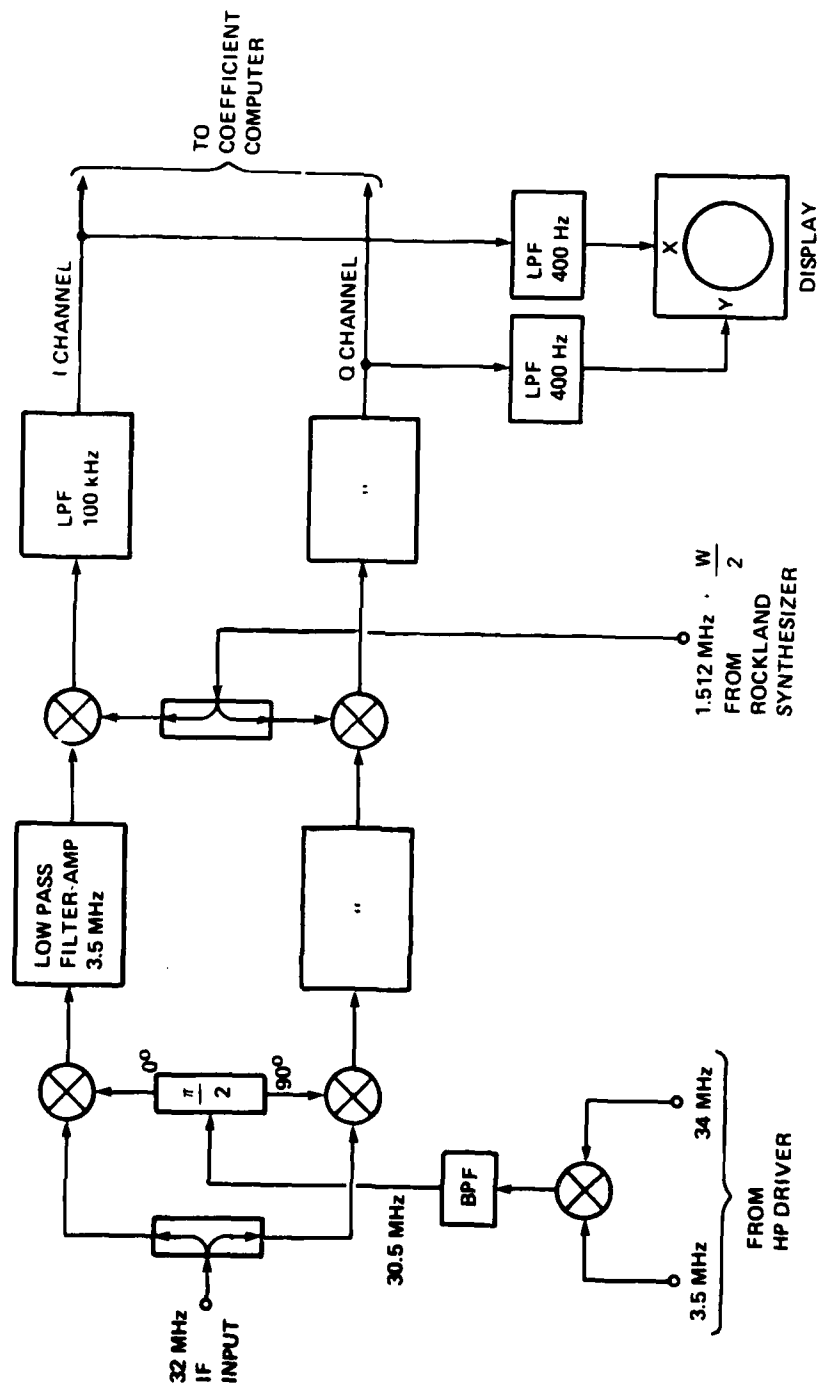


Figure 34: RECEIVER BACK-END AND DISPLAY

I and Q signals are sent (1) directly to the coefficient computer's ADC's and (2) via 400 Hz LPF's to a CRT display which is used for signal acquisition (as described in sections 3.2.1 and 3.5.1).

#### 4.5 Spectrum Analysis Equipment

To spectrum analyze the interference in either a 1024-kHz or 512-kHz band, the Rockland synthesizer is programmed to output an LFM waveform which sweeps over the desired band in one second. The Rockland output is up-converted to 32 MHz where it is used to synchronously detect a 32-MHz IF amplifier output. A 3-kHz LPF sets the predetection bandwidth to 6 kHz. At each site the LPF output is recorded on an audio cassette tape recorder along with 1 Hertz timing marks. Later both cassettes are played back at MITRE/Bedford, and the signals are detected, log-amplified, and recorded on an oscillographic chart recorder.

The spectrum analysis circuitry in the receive terminal is shown in Figure 35. In the transmit terminal, the exciter is reconfigured as a receiver whose output is filtered in a 32-MHz IF amplifier and fed to the spectrum analysis equipment. Figure 36 shows this arrangement with those circuits not used in the exciter enclosed by a dashed line. (This equipment is not shown in the test facility system block diagram of Figure 16).

#### 4.6 Peripheral Equipment

Equipments for voice communications, display, and recording not previously described are discussed in this section.

##### 4.6.1 WWV/CHU Reception and Synchronization

A McKay-Dymec DR55 Communication Receiver is included in each terminal. This equipment receives broadcasts from WWV or CHU, usually via a whip antenna, and outputs a one-Hertz time "tick". This signal

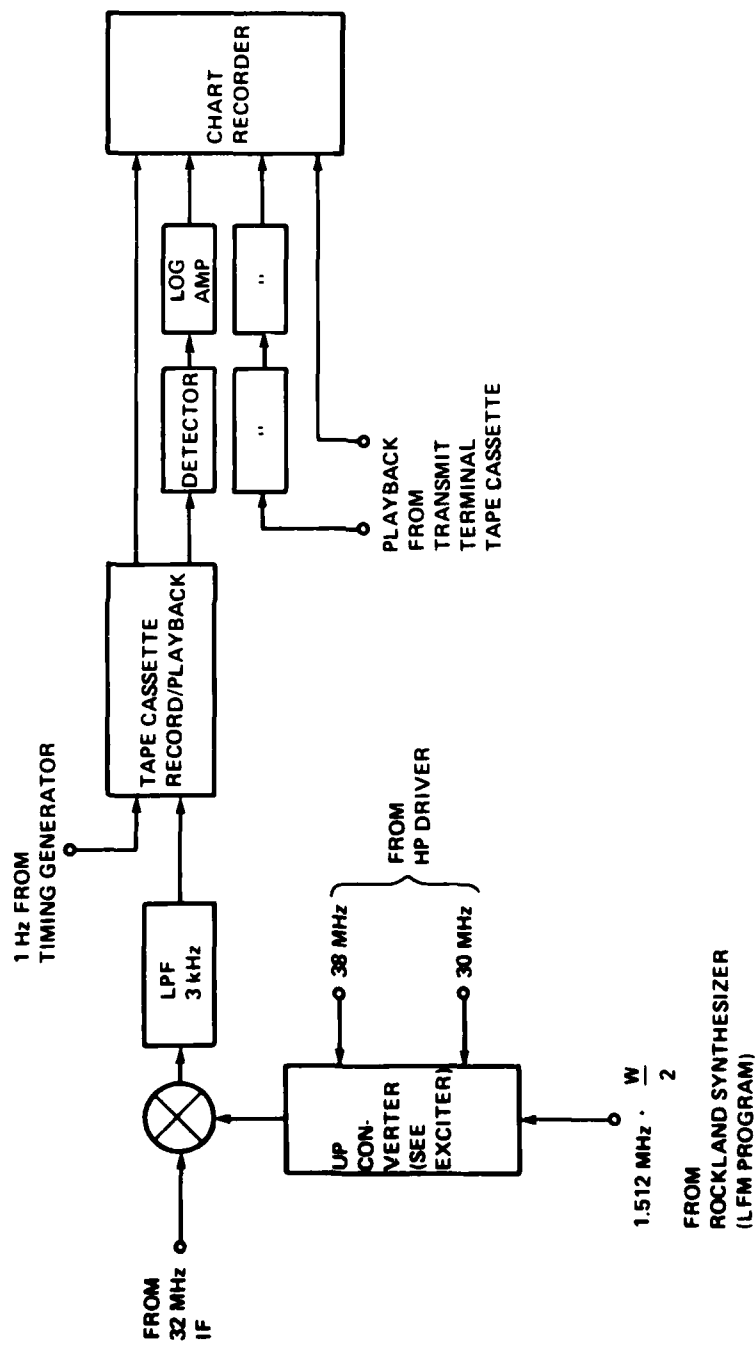


Figure 35: RECEIVER TERMINAL INTERFERENCE SPECTRUM ANALYSIS EQUIPMENT

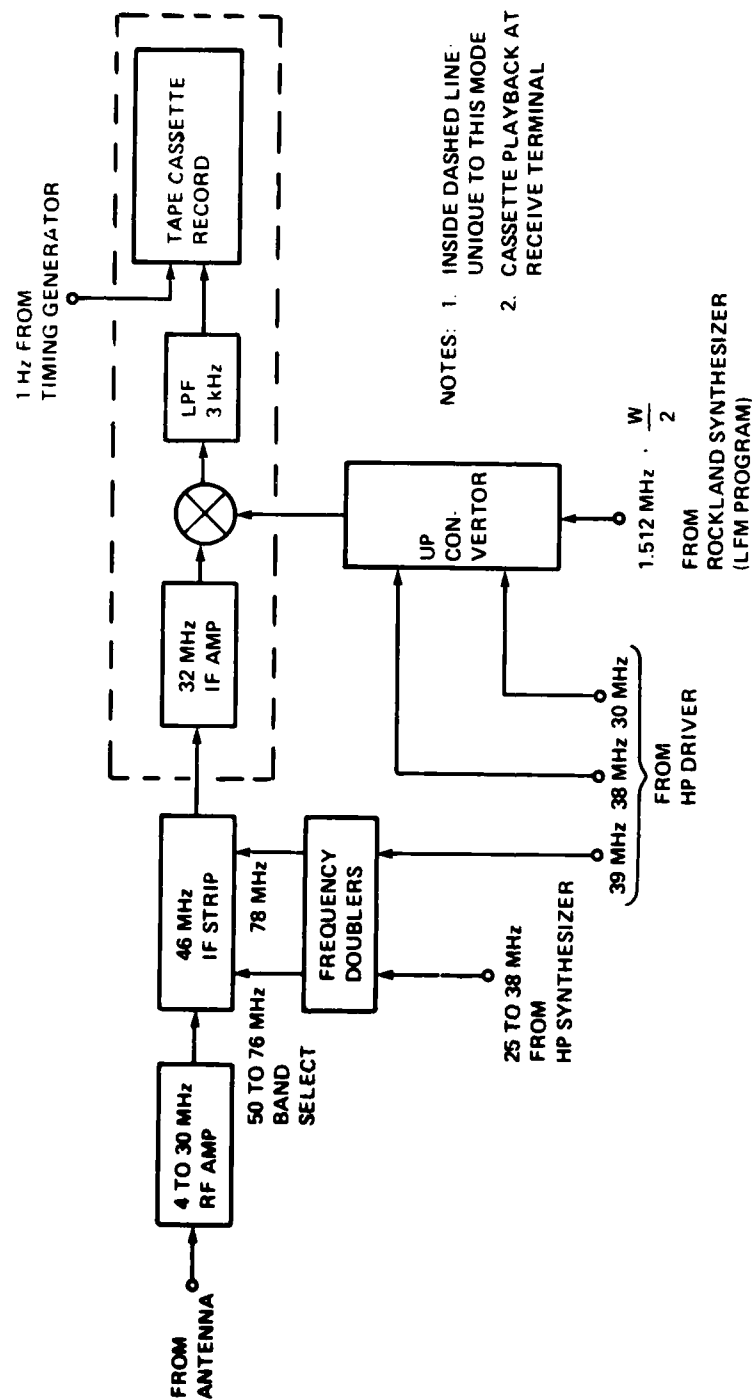


Figure 36: TRANSMIT TERMINAL RECONFIGURED FOR INTERFERENCE ANALYSIS

is compared to the local one-Hertz time tick from the timing generator on an oscilloscope, and the local clock is interrupted for the appropriate time duration so as to achieve coarse epoch synchronization at each terminal. The receiver terminal clock is next interrupted for the estimated propagation delay between the terminals, and then the receiver back-end display is used for signal acquisition as described in sections 3.2.1 and 3.5.1. This equipment is shown in Figure 37. (It is not shown in the test facility system block diagram of Figure 16.)

#### 4.6.2 SSB Voice Communications

Collins KWM-2A 3.5- to 30-MHz transceivers with 100-watt output capability are used for voice communications between the transmit and receive terminals. Appropriate HF channels have been allocated, and either a whip or a LPA antenna is used at each end of the circuit.

#### 4.6.3 Wideband Recording Facility

The Bell and Howell VR-3700 Tape Recorder is a 14-channel machine. It is operated at 60 inches per second where it provides a 0.4- to 750-kHz channel bandwidth.\* Three recording tracks are made on each tape. Each track uses four channels: (1) I channel, (2) Q channel, (3) time reference, and (4) frequency reference for playback speed control. The phase error between the I and Q channels has been measured as  $20^{\circ}$  RMS for the worst track.

The I and Q channels are used to record the input to the equalizer. Playback is either through the equalizer or around it to the Interface Unit's up-converter. This equipment is shown in Figure 38.

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\*The information below 0.4 kHz is lost but this only comprises 0.004 or 0.002 of the total signal spectrum for  $W = 512$  kHz and 1024 kHz respectively.

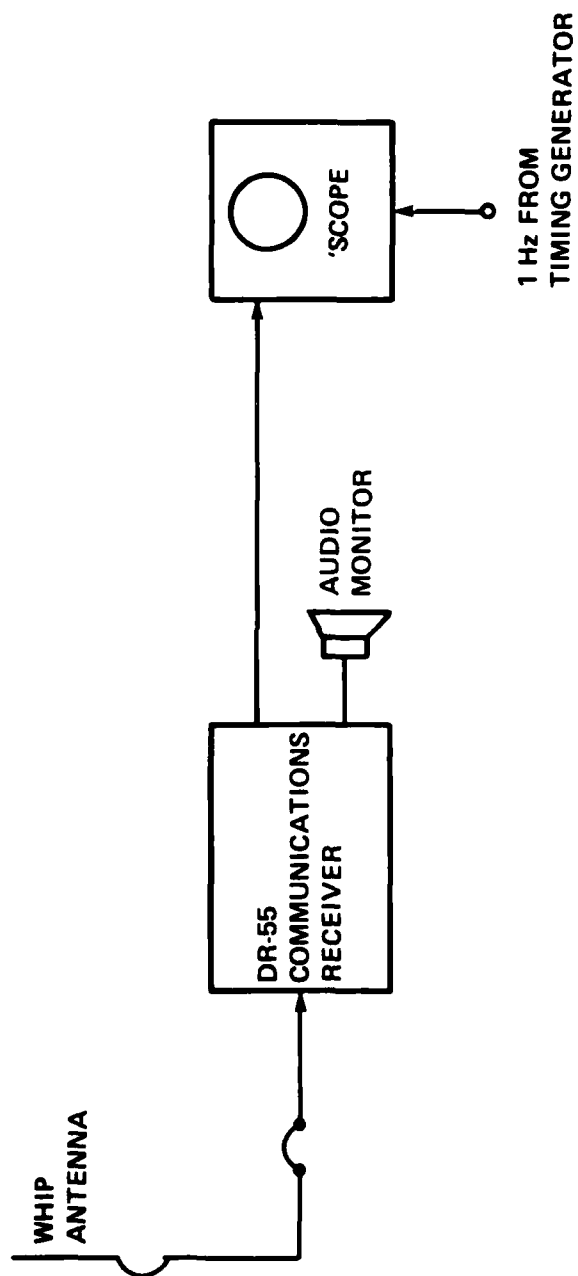


Figure 37: WWV/CHU RECEPTION AND SYNCHRONIZATION

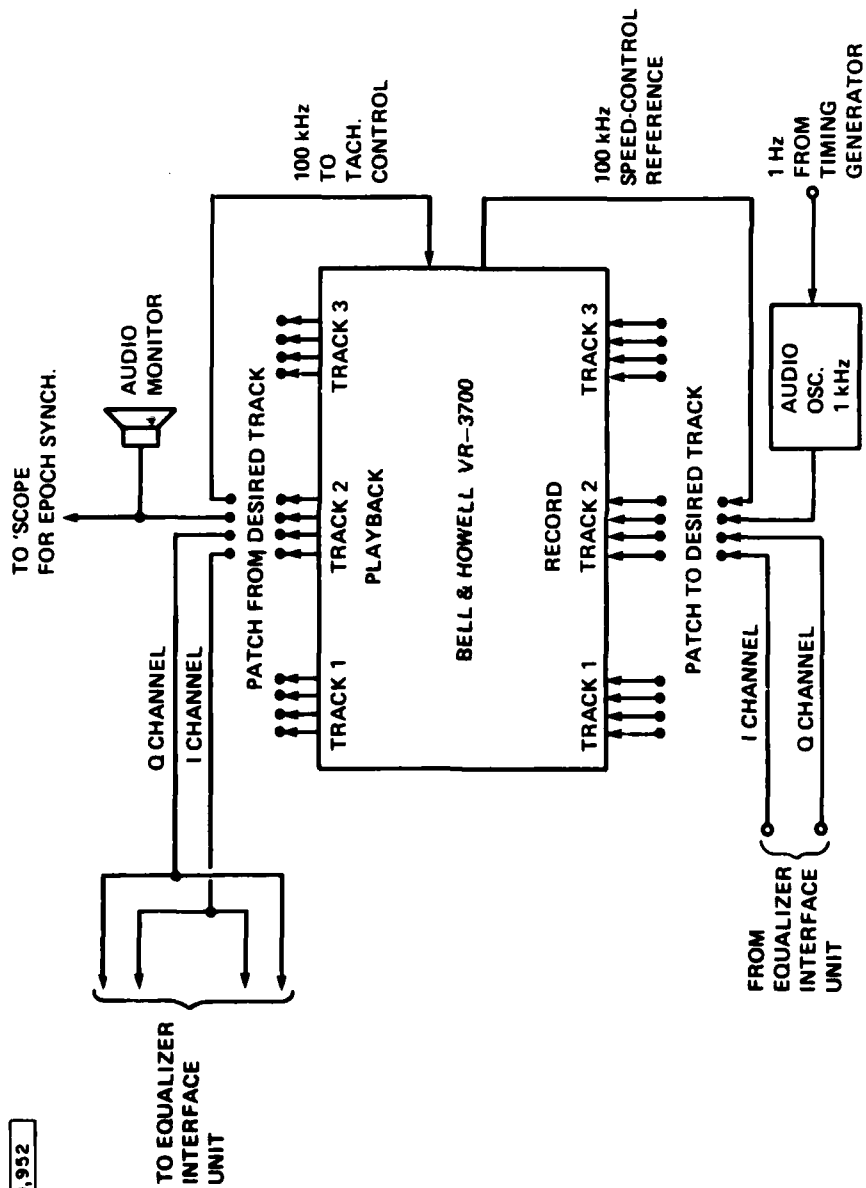


Figure 38: WIDEBAND RECORDING FACILITY



#### 4.7 PN Spread Spectrum Capability

A discrete sequence PN phase modulation and demodulation capability is not presently available but will be added to the test facility since it appears to have advantages for low-probability-of-intercept communications which complement the advantage the FFH waveform has with respect to anti-jamming.

The PN waveform will be a simple bi-phase modulated signal at a chipping rate of 1024 or 512 kb/s. A fairly long repeating linear sequence with good autocorrelation properties can be used to permit mode isolation in the receiver signal processing. This can be accomplished by limiting the time duration of the receiver correlator window to less than the typical mode spacing. For convenience the system will use a 125  $\mu$ s correlation window.

To process the PN signal in the receiver, one alternative is to use the equalizer to also serve as a matched-filter correlator by loading the appropriate N chips (usually 128) from the receiver PN sequence generator into the most significant bits of the A & B registers every T sec (usually 125  $\mu$ s). One-bit multiplication (exclusive - OR) of the PN code and the coefficient computer output is required. Another alternative is to implement separate circuitry devoted to the correlation task. Charge transfer device correlators are available which are suitable for this application.

## SECTION V

### EXPERIMENTAL PROGRAM

Preliminary (shake-down) system tests have already been completed (May 1980) between Boston Hill and MITRE/Bedford. In these tests, the transmit terminal and the equipment in the receive terminal required for probe signal reception and acquisition were successfully checked out.

The complete test facility as described in section IV will be first used in a proof-of-concept demonstration on a Florida-to-MITRE/Bedford ionospheric path. This demonstration will also be the beginning of an extensive experimental measurement program to establish a data base for future wideband HF modem designs. The transmit terminal will be moved to several points in North America and signals received from the transmit terminal at MITRE/Bedford will be recorded and analyzed.

The available choices of spread spectrum modulation will be used and the equalizer will be configured in various ways using the built-in flexibility with respect to W, T, and its dynamic range. The interference spectrum will also be recorded at both sites.

The planned upgrading of the test facility to include a PN spread spectrum capability (see section 4.7) will proceed in parallel with the measurement program. When this upgrading is completed, additional testing from a site (or sites) already used will be undertaken.

#### 5.1 Data Taking

The following ionospheric paths will be investigated:

- short range (20 km): from MITRE's facility at Boston Hill in North Andover, MA.

- long range, one-hop (1.5 to 2 megameters) from the south-eastern CONUS,
- long range, multihop (3 to 4 megameters) from the western CONUS, and
- near the auroral oval, from Newfoundland.

Results from subsequent analysis (see section 5.2) will be used to refine the type of data taken as the program progresses. It may be desirable to repeat some of the early measurements to incorporate these refinements on a comparative basis and to investigate the ionospheric variation over a transseasonal interval.

Wideband recordings of the receiver output (unequalized) will permit later playback via the equalizer during which the equalizer parameters can be varied.

## 5.2 Data Analysis

The data base to be established can be described as a matrix of data sets having the following four dimensions:

- 1) Distance and/or location: Path lengths from 20 km to over 3000 km oriented south-north and west-east in the temperate zone and also a path near the auroral oval.
- 2) Time: Diurnal (daytime, dusk, nighttime) and seasonal.
- 3) Spread spectrum waveforms: FFH and PN, 1024 and 512 kHz bandwidths.
- 4) Equalizer parameters: 64 and 128 taps, up to 8 bits of amplitude, 1024 and 512 kHz bandwidth.

These data will be analyzed to determine quantitative relationships between the various dimensions. For example, assuming 24-hour operation, recommended equalizer parameters can be derived for a particular application as described by dimensions (1) and (3).

The interference power spectrum data will be analyzed to determine the correlation of the measurements at the transmitter site with those at the receiver site and to derive recommendations leading towards equipment or techniques to mitigate the effects of this interference.

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